VHF HANDBOOK

VERY HIGH FREQUENCY

BY WILLIAM I. ORR - W6SAI
AND H. G. JOHNSON - W6QKI

$2.95

VHF

Very High Frequency

HANDBOOK

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Herbert G. Johnson, W6QKI

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THE VHF HANDBOOK

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FOREWORD

VHF is the wave of the future. Discovered by Hertz, pioneered by Marconi, and utilized by radio amateurs for over a quarter-century, the Very High Frequency portion of the radio spectrum promises tremendous potential for the future. The recent electrifying experiments culminating in reliable beyond-horizon VHF transmission has unchained this range of frequencies from line of sight paths and have opened a new communication frontier.

The fact that VHF offers the only consistent form of beyond-horizon propagation gives pause for thought. Unaffected by the vagaries of the ionosphere that torment and disrupt long distance communication on the high and medium frequencies, the broad region above 50 mc offers unparalleled advantages to all modes of international and intercontinental communications. It is not unreasonable to imagine international television transmission by means of VHF scatter links, or trans-ocean telephone and teletype circuits operating in the VHF range.

Although the basic VHF knowledge is years old, this portion of the radio spectrum is yet in the morning of its youth. Many wonders are still to be unveiled in this exciting frequency range. Much that we now accept for fact will be supplanted with newer knowledge. Much that we suspect as fancy will be commonplace. The newest designs and ideas of today will become obsolete tomorrow.

As in the past, the radio amateur will be in the forefront of the coming developments and discoveries in this fascinating and vital field. To these pioneers in the frontier of the future, this book is dedicated.

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CHAPTER I

The VHF Region

The curtain was lifted on the field of Very High Frequency radio transmission on a cold, blustery spring day in the year 1884. At the University of Kiel in Germany young Heinrich Hertz observed transmission of electromagnetic energy between two resonant spiral coils, one of which was attached to a charged Leyden jar. This history making electronic discharge took place in the region of the present two meter band. Later experiments by Hertz and Ruhmkorff established resonant transmitting and receiving systems for the transfer of this new, invisible form of radiant energy. The distances covered were minute; the equipment crude. Little did the early pioneers imagine the vast and fertile territory at whose frontiers their experiments were taking place. It is interesting to note that the first attempts at electromagnetic communication took place in the VHF region, but for years afterwards this territory was forgotten. The experiments of Hertz gathered dust in archives as the efforts of the radio pioneers followed other phenomena in the low frequency fields.

Nearly half a century later in 1932 Marconi established communication

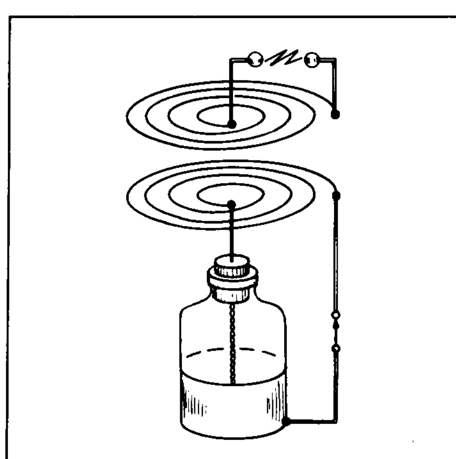


Fig. 1 The simple apparatus of experimenter Heinrich Hertz produced damped VHF oscillations in the vicinity of the present 2 meter band. Discharge of energy in lower coil produced spark across gap of upper resonant coil.

over a distance of 168 miles on a frequency of about 500 megacycles, and predicted that extremely short radio waves could be propagated great distances beyond the optical horizon.

Nevertheless, the interest of the radio engineers and amateurs were fixed in the 400 to 10,000 meter region of ground wave communication. (In those early days, "radio waves" were generally measured in terms of wavelength, with the Metric System being universally adopted as the standard system of measure.) The relatively simple equipment at hand readily generated and detected waves in this region, but equipment for work at short wavelengths was unknown. At wavelengths below 200 meters the groundwave rapidly expired, and it was deduced that the path of such waves would end abruptly at the optical horizon. These commercially useless frequencies were turned over to the radio amateurs and experimenters for their use.

Ionospheric Reflection

The discovery of ionospheric reflection of wavelengths in the 10 to 100 meter region brought a rush of activity to this portion of the spectrum. Great hope was held that as the frequency of propagation was raised the ionospheric skip-distance would increase. When the waves below 8 or 10 meters seemed to revert back to optical paths and showed no signs of ionospheric reflection, interest in this portion of the radio spectrum waned.

A few experimenters, however, still worked at the problem of generating and detecting the higher radio frequencies. In 1920 Doctors Birkhausen and Kurz generated extremely short radio waves with positive grid vacuum tube oscillators. Marconi conducted VHF tests between land and his yacht *Electra,* and noted that these radio signals often provided communication well beyond the line of sight, indicating that these tiny waves were not chained to the optical horizon. In the year 1924, independent observers in several countries noted that VHF waves could be reflected from ships and aircraft. Secret experiments along this line led to the development of *radar* (radio detection and ranging) which played so prominent a part in World War II. All tests and experiments, however, pointed to the horizon as a practical limit of VHF radio propagation. The isolated cases of beyond-horizon propagation were assumed to be due to unexplained vagaries in the transmission path.

Early Amateur VHF Work

During the eventful year of 1924 the first amateur experiments took place in the vicinity of 6 meters. Radio 1FG and radio 9APW conducted experimental transmissions at this wavelength, using de-based UV-202 oscillator tubes. In 1925 9EK experimented with transmissions on 4 meters, and an exclusive amateur frequency assignment of 400-400.1 mc was made by the U. S. Department of Commerce.

A portent of things to come were the c-w tests conducted by 9EHT in 1926. His 56 mc signals were heard in both Connecticut and New York, a distance of over 1000 miles. The next year brought increased interest in the amateur 56 mc band, and a classic article in *QST* by George Hart showed the experimenter how to build a simple 56 mc super-regenerative receiver. By the year 1930 considerable activity was in evidence on this band, most stations employing modulated oscillators and super-regenerative receivers.

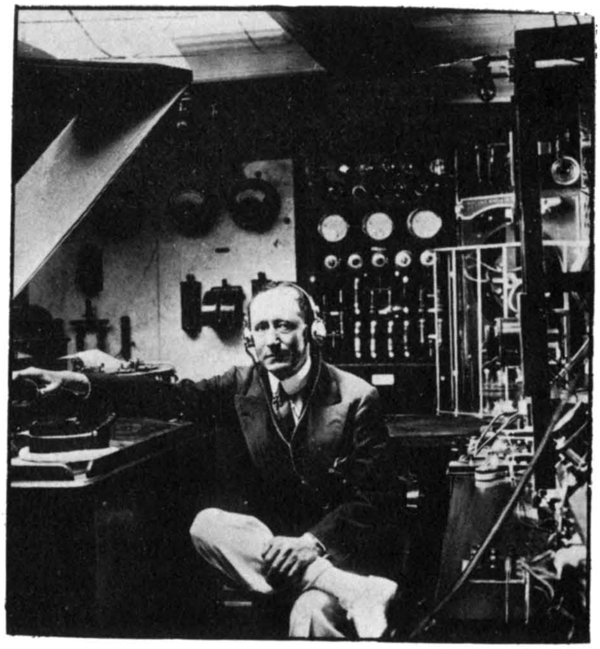


Fig. 2 "The Father of Radio," Guglielmo Marconi experimented with centimeter waves, sending messages on 500 mc between his yacht and the land over a distance of 168 miles.

Propagation, in general, was confined to line of sight transmission paths. Long distance contacts were limited by the simple receiving and transmitting equipment then in use. The VHF horizon for consistent communication seemed to extend to about 1.3 times the optical horizon, and the few contacts obtained past this range were intermittent and decidedly unreliable.

The early "thirties", however, mark the beginning of VHF communication as we know it today. Multi-element parasitic arrays were used on 150 mc by W6AJF in 1933, and about this time the famous 5 meter "transceiver" was born. The eternal bane of the serious VHF operator, these simple two tube stations gave a tremendous impetus to high frequency operation by radio amateurs. Developments in amateur activity on the higher frequencies began to quicken and in 1935 the amateur world was electrified by the report that the 56 mc signal of W2DEE had been heard in Michigan. Other reports of long distance reception were recorded, and on June 22, 1935 W1CBJ had a verified 56 mc contact with W8CYE over a 900 mile path. During the next year or so sufficient instances of two way VHF communication were noted in the 56 mc band to establish the idea that sporadic communication could be supported on frequencies appreciably above those normally reflected from the ionosphere. At the same time, the state of the art advanced rapidly, as VHF interest grew. By 1937 superhetrodyne receivers and crystal controlled transmitters were in use, and in 1939 W6GPY and W6ZA had two way communication on 235 mc.

The VHF aspects of amateur radio had come a long way from the first tentative experiments of Hertz. By the middle "thirties" the use of frequencies up to 30 mc or so was quite clearly understood. In general, the spectrum above this point was still considered to be of little value, the few mobile services near 35 mc and early FM and TV broadcasts in the 45 mc region were merely indications of things to come. The intermittent reports of beyond-horizon communication on the higher frequencies seemed to be too erratic and spotty to be of much practical interest.

The Smooth Sphere Theory

In 1937 Van der Pol and Bremmer published the classic *smooth sphere* theory of VHF propagation. According to this theory, VHF signals would experience an exponential drop in strength beyond the horizon of about 1.2 db per mile at a frequency of 500 mc, and a drop of about 2.4 db per mile at 4000 mc. This theory was based upon the assumption that the earth was a smooth sphere, that the composition of the atmosphere was uniform, and that the effective "radio radius" of the earth was 4/3 of the true radius. The last assumption took into effect the atmospheric refraction of VHF signals that had been observed.

The smooth sphere theory held true for signals at short distances beyond the horizon, but instances of isolated cases noted during 1936-1940 raised serious questions that could not be answered by this theory. It was obvious that intermittent forces were at work to extend the range of VHF signals at random times and locations.

Just before World War II, certain controlled experiments began to show promise of semi-reliable communication beyond the line of sight on frequencies well above those normally reflected back to earth by the ionosphere. Prominent among these experimenters were many radio amateurs, the most noted being the late Ross Hull, W1AL. Using high gain antennas, powerful transmitters and sensitive receivers, reliable communication was maintained on 112 mc on a beyond-horizon path. It was noted that such signals faded over a wide range of amplitude, and that signals at this great distance were of greater strength than predicted by the classic smooth sphere theory. A concept of atmospheric ducts and temperature inversions was introduced to explain these unusual propagation ranges.

By the eve of World War II a large file of information concerning beyond-horizon communication in the VHF region had been accumulated. In general, early DX efforts in the 56 mc band depended upon the sporadic capacity of the ionosphere to reflect unusually high frequencies, the most noted effects being *aurora propagation* and *sporadic E-layer skip*. Distances of several hundred miles were covered in the 2½-and 1¼-meter bands as a result of anomalities in the atmosphere, principally temperature inversions. There was no consistency to the reports of communication, and insufficient information was at hand to formulate a new theory of VHF long distance transmission. Rather, the situation was looked upon as a discovery of intermittent forms of long distance interference to already established line of sight VHF communication links. Beyond providing the amateur with thrilling moments of VHF DX, the sporadic periods of beyond-horizon propagation held little promise of being more than a curiosity.

Two technical break-throughs occured just before the War that promised great achievements in the VHF region. In 1938 the Varian brothers developed the Klystron tube capable of generating large amounts of power well into the centimeter region. In England, in the same year a research group at the University of Birmingham developed the resonant cavity magnetron, which was capable of great power output at extremely short wavelengths. As a final touch, this period saw the first transcontinental F_2-propagation contact on 56 mc between W1EYM and W6DNS as the sunspot cycle reached a new peak.

The Post-War Period

During the War there became available to the government and commercial investigators transmitting tubes of fabulously greater power than had ever been produced. High gain radar antennas and low-noise receivers were developed, and untold dollars and man-hours were spent in the investigation of propagation phenomena for military purposes in the VHF and UHF regions. Radar ranges greatly in excess of those predicted were noted, the most famous observation being the radar reflection of the Arabian peninsula observed from India during the closing stages of the War. VHF point to point circuits were producing greater ranges than those normally expected by application of classic theory. Ionospheric stations were examining the great reflecting mirror, and the science of predicting MUF (maximum useable frequency) and radio propagation came into being.

Immediately following the close of the War numerous tests were conducted in the region of 50 to 100 mc, and special hearings were held before the Federal Communications Commission to determine the true state of VHF propagation. One of the results of these hearings was the temporary "freeze" placed upon the construction of new television stations, brought about because the amount of co-channel interference proved to be much greater than predicted by the smooth sphere theory of VHF propagation.

The mass of accumulated data pointed to the fact that beyond-horizon signals proved to be much stronger than theorized, and that frequencies in the range of 30 to 50 mc could be consistently received over paths up to 300 miles in length when high radiated power and sensitive receivers were employed. Similar results had been observed at frequencies well up into the microwave region.

The increased interest of commercial and government facilities in the development of the VHF region meant that much of the tests and experimental work would be done by controlled laboratory experiments, rather

Fig. 3 Using a 20 kilowatt, 418 mc Resnatron tube, Collins Radio Co. sent a coded message from Cedar Rapids, Iowa, to Sterling, Va. via moon reflection.

than in ham "shacks." However, the amateurs played an important role in these developments, as many persons of prominence in these research programs were themselves amateurs. In addition, many instances of amateur VHF development and contribution were to be noted.

Aurora contacts on the 56 mc band were first noticed by amateurs in the early "thirties," and at a somewhat later period on 144 mc. During 1953 many amateurs cooperated with the Cornell University Ionospheric project by making observations and reporting aurora contacts to the University. In addition, W2MTU made significant contributions to the theory of this type of VHF propagation with his studies at the University of Alaska in 1953.

An amateur meteor reflection link was established between W4HHK and W2UK which ran for many months during 1953. The recorded results of these 900 mile tests on the 144 mc band aroused great interest in both amateur and scientific circles. W2NLY, W2AZL, and others also joined in these tests. Later, in 1955 W4HHK also conducted meteor reflection tests with W1HDQ over a 1020 mile path.

Starting in 1949, the Radio Amateur Scientific Observation Project (RASO) sponsored by the U.S. Air Force and *CQ* magazine and directed by O. P. Ferrell was organized to collect data on the prevalence, distribution and intensity of signals propagated by sporadic-E ionization on the 6 meter band. Hundreds of amateurs contributed to the work of this project.

As a result of combined attacks upon the field of beyond-horizon VHF transmission the properties of the frequency spectrum above 30 mc were slowly revealed. Even though these various modes of propagation produced astonishingly strong VHF signals at distant points, the transmission path seemed to be exceedingly unreliable, and was dependent upon a fortituous combination of natural circumstances beyond the control of man. If the right combination was in evidence, the path could be used. But no means were at hand to allow the VHF operator to control the availability of the path.

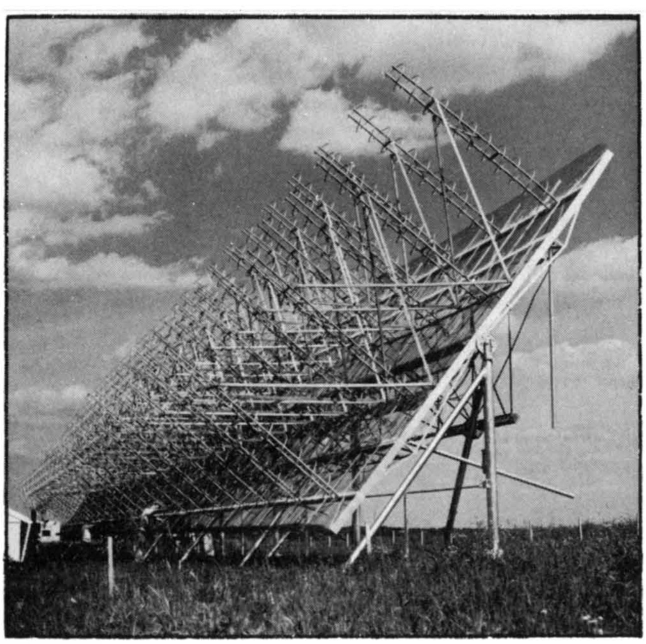


Fig. 4 Ninety-six element beam used by Dr. Kraus (W8JK) at Ohio State University. Helical arrays are used to chart radio signals received from outer space. One degree beam pattern permits charts to be made of VHF radiation received from various sections of the universe.

Scatter Signals

The cumulative results of years of efforts have shown that there are weak non-optical VHF signals which are everpresent, and which do not depend upon anomalies or intermittent atmospheric phenomena for their existance. These signals have been termed *scatter signals.* The actual discovery of scatter-type VHF signals was made by many independent individuals and organizations at about the same time, since most of the observed data pointed to such propagation phenomena. The first announcement the general body of amateurs had of tests being conducted with scatter signals was the impact of the 49.8 mc test signal radiated by the Collins Radio Co. 20-kw "scatter" transmitter in January, 1951. The "big Iowa signal" was heard up and down the eastern seaboard, at a distance of over 770 miles. The National Bureau of Standards receiving station at Sterling, Va. continuously monitored the test signal, and noted that there was a minimum level beneath which the signal would not fall, and that this minimum level signal was ever-present, regardless of ionospheric conditions. Similar test transmissions were made on a frequency of 107.8 mc. These signals did not depend upon anomalies or intermittent phenomena for their existance. They provided reliable communication on a beyond-horizon path 24 hours a day, something that was impossible on any lower frequency!

A significant contribution to the scatter technique investigation was made in 1951 when WϕCXX maintained a scatter propagation schedule with an amateur station in Dallas, Texas on 29.7 mc. This non-optical path provided reliable communication at any time of day or night, regardless of whether the 10 meter band was "open" or not.

Mike Villard, Jr., W6QYT of Stanford University made a unique and valuable contribution to the scatter study in 1951 when he investigated the properties of back scatter of ionospheric-reflected signals. Scatter echoes received on the 14 mc band were observed on a radar scope, which gave a picture of actual ionospheric openings on the band as shown by echoes received from transmissions at the frequency. This technique allowed examination of a certain transmission path without the necessity of two way communication over the path. Also prominent in this study was R. A. Helliwell, W6MQG, director of Ionospheric Research at Stanford University.

A summary of VHF propagation was made in 1953 by M. G. Morgan, W1HDA, Chairman of the IRE Subcommittee on Ionospheric Propagation. It was now apparent that a new means of propagation of unheard-of reliability was at hand for medium distance communication. The economic and military possibilities of such a discovery are of far reaching consequences, and great effort is being expended in further research into this important aspect of VHF communication.

In summary, it can be stated that there are two broad forms of beyond-horizon propagation at work in the VHF spectrum. The first of these groups may be considered as dependent upon atmospheric anomalies or irregularities for its existence. Included in this group are Sporadic-E, aurora, air-mass boundary bending, F_2-layer reflection, meteor trail reflection, and others. The second group encompasses the scatter signals. These signals do not depend upon such atmospheric irregularities. The scatter signals are much weaker than those listed in the first group, but the latter are completely independent of the vagaries of propagation that plague long distance trans-

mission. Atmospheric irregularities, sun-spot storms, radio black-outs and other disturbing effects have little effect upon the scatter signal.

The irregular properties of the first mentioned group of propagation mechanisms have long been the delight of the ham and the source of interference to the commercial communication circuits operating in the VHF region. The reliability of the newly discovered scatter-type propagation opens large and expanding fields for the future.

Beyond-Ionosphere Signals

An intriguing aspect of VHF communications is that the frequencies above the MUF pass with ease through the ionosphere. By the end of World War II magnetrons with peak powers of more than a million watts, and resnatrons with continuous power capabilities as great as 50 kw were at hand. Computations indicated that within the power limit available it would be possible to bounce the ionospheric-penetrating VHF waves off the surface of the moon, and to receive the return echo on earth. This feat was accomplished by Lt. Col. John De Witt, Jr., W4ERI of the U.S. Army Signal Corps in 1946 using a modified 8 kw radar set operating on 111.5 mc. An amateur version of the Signal Corps' "Project Diana" was started in early 1950 by W4AO and W3GKP. Tentative moon reflections of the 1 kw, 144 mc transmissions were noted in the summer of 1950, and in the spring of 1953 W3GKP and W3LZD obtained moon-reflected echoes of the W4AO transmissions. In September, 1953, Herbert Johnson, W3QKI (W6QKI) received his own echoes from the moon, using a 104 element beam antenna and a 700 watt transmitter. A few months earlier, the Collins Radio Co. established a moon-reflection link between Iowa and Washington, D.C., using a frequency of about 400 mc.

It is in this VHF region that future communication between the earth and unmanned space satellites will take place during the span of the next few years. The compact VHF transmitters and small, efficient radiating systems, combined with the ability of these waves to puncture the ionosphere make the VHF range of frequencies the communication bridge between the earth and other planets.

A pioneer in the field of interstellar radio signals is Dr. John Kraus, W8JK, of Ohio State University who has been studying the radio emanations of distant stars with the use of a giant "radio telescope." Experiments along this line are also being conducted in England, with the use of sensitive receivers and gigantic directional antennas. Thus the properties of the universe about us are being investigated by the reception of distant electromagnetic emanations from other solar systems. In addition, it has recently been noticed that other planets of our own solar system are radiating radio waves that are providing useful information for us.

The VHF Communication Circuit

Regardless of the mode of propagation of the VHF signal, the communication circuit over which it passes may be broken down into separate functions for examination and study. Although this text is primarily concerned with the transmission and reception of voice-modulated and c-w signals, particularly within the amateur bands, much of the information

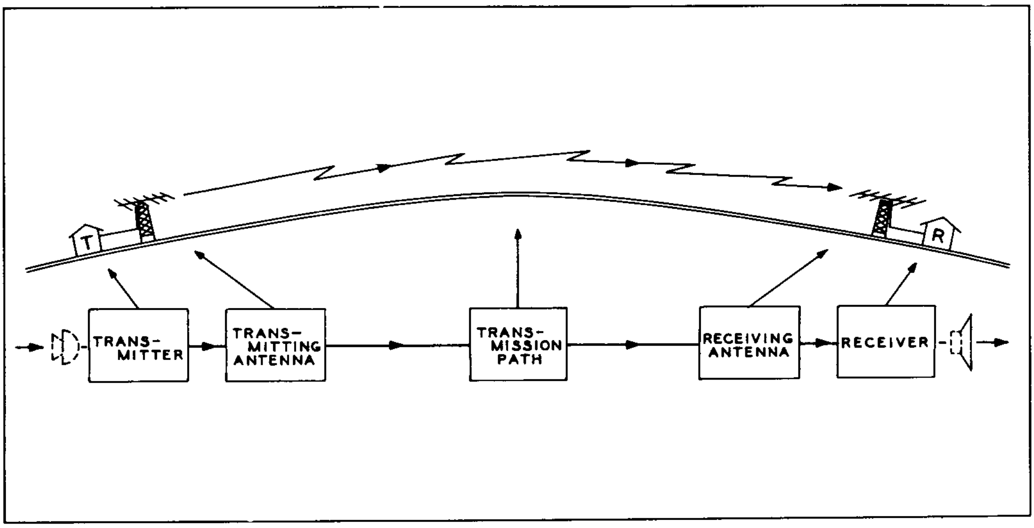


Fig. 5 The VHF communication circuit is composed of five important items, as shown above. Each item contributes toward the operation of the circuit, and each affects the efficiency of the circuit, as discussed in the text.

herein applies equally well to commercial circuits, such as FM, TV, and the mobile services. All of these circuits exist for the purpose of transmitting intelligence from one point to another.

A typical VHF communications circuit is pictured in Figure 5. The essential components of this circuit are: 1—The r-f generator (which includes a form of modulation, or a keying system). 2—The radiating system at the transmission end of the path. 3—The transmission path. 4—The antenna system at the receiver, and 5—The receiving equipment, which includes a suitable demodulating system to recover the transmitted intelligence. Each of these components is important, and the operation of each affects the overall efficiency of the circuit. These components will be considered at length in this Handbook.

Of great interest to the VHF engineer are the various factors which influence the efficiency of the communication circuit. The most important of these factors is the *background noise level.* This is the residual noise in the system with which the received signal must compete in order to be useful. The desired quality of reproduction determines how much stronger than the noise level the signal must be. The familiar term *signal-to-noise (S/N) ratio* is used as a measure of how many decibels the signal rides above the noise level. In very minimum circuits, such as amateur communications where weak signals are acceptable and reliability is not so important, it is possible to work with S/N ratios as low as -3 db or less. That is, the signal is actually below the noise level. On the other hand, in certain commercial circuits it is not uncommon to operate a circuit with a S/N ratio of plus 60 db or more. Naturally, the power and transmission path requirements in this latter case become more demanding than in amateur communications.

Background noise level manifests itself in the receiving equipment as a part of the detector output, and is usually a random series of impulses scattered across that portion of the frequency spectrum passed by the detector. Audibly it is identified as a "hiss" or rushing sound. Visually, as on the screen of a television receiver, it shows up as "snow," or a microbe-like movement over the picture.

Circuit Noise Sources

Each component of the communication circuit contributes a share of noise to the final output noise observed at the receiver. Usually the majority of noise comes from one or more of the following sources: 1—*Shot effect* in the early amplifier stages of the receiver. 2—*Thermal agitation* generated in the antenna system. 3—Static discharges in the atmosphere, and 4—Man-made interference.

At lower frequencies the shot noise generated by the vacuum tube amplifiers of the receiver is quite low, and most of the noise output of a sensitive receiver can be attributed to thermal-agitation or *Johnson Noise* from the antenna, plus atmospheric static. However, as we move above 50 Mc, tube noises become more noticeable and receiver circuits and vacuum tubes must be specially designed to minimize such factors. Above 100 Mc it is very important to choose a "low noise" type tube for the first and second r-f amplifier stages of the receiver, since these stages will largely determine the overall *noise figure* of the receiver. (This term is a figure of merit which evaluates VHF receiver performance, and represents the ratio between the noise generated by a perfect receiver, and the noise generated by the receiver under test.) These sources of circuit noise will be discussed at length later in this Handbook.

A second factor which influences the overall efficiency of the communication circuit is the segment of the spectrum that is allowed to pass the output terminals of the receiver. This segment is defined as the *receiver bandwidth.* It is highly important that this segment be kept to the minimum necessary amount, since S/N ratio is directly affected by the width of the receiver passband. This is true since the external noise level is received over a wider portion of the spectrum in a receiver of broad bandwidth, and the noise output of the receiver is in direct proportion to the bandwith. The signal output of the receiver does not increase as the passband is increased past the minimum necessary amount. For example, in voice communications it has been found that the audio frequencies in the range of 300 to 3000 cycles transmit the greater portion of sounds necessary for good intelligence. These sounds lie in a bandwidth of 2700 cycles, referred to as the *optimum bandwidth* for voice reception. Any bandwidth greater than this figure will not materially improve the intelligence, but will detract from the S/N ratio, thus placing a greater burden on some other portion of the communication circuit. If the receiver bandwidth is twice as wide as necessary, the S/N ratio is degraded by 3 db.

For c-w reception it is possible to reduce bandwidth to a very narrow spectrum. The *slower* the rate of information transmitted, the narrower may be the bandwidth alloted to transmission of the intelligence. Simple c-w systems have a minimum practical bandwidth of about 100 cycles.

A third factor which influences the overall efficiency of the communication circuit is the *transmitter power* which may be thought of as depending upon such variables as receiver bandwidth, receiver noise figure, antenna gain at each end of the system, desired S/N ratio, and the attenuation of the transmission path. The latter is the most difficult variable to predict, and usually can only be determined by a series of tests. If the path experiences periodic variations in signal strength, then the desired percentage of circuit reliability must also be considered. Greater transmitter power is necessary to fill in the periods of higher signal attenuation.

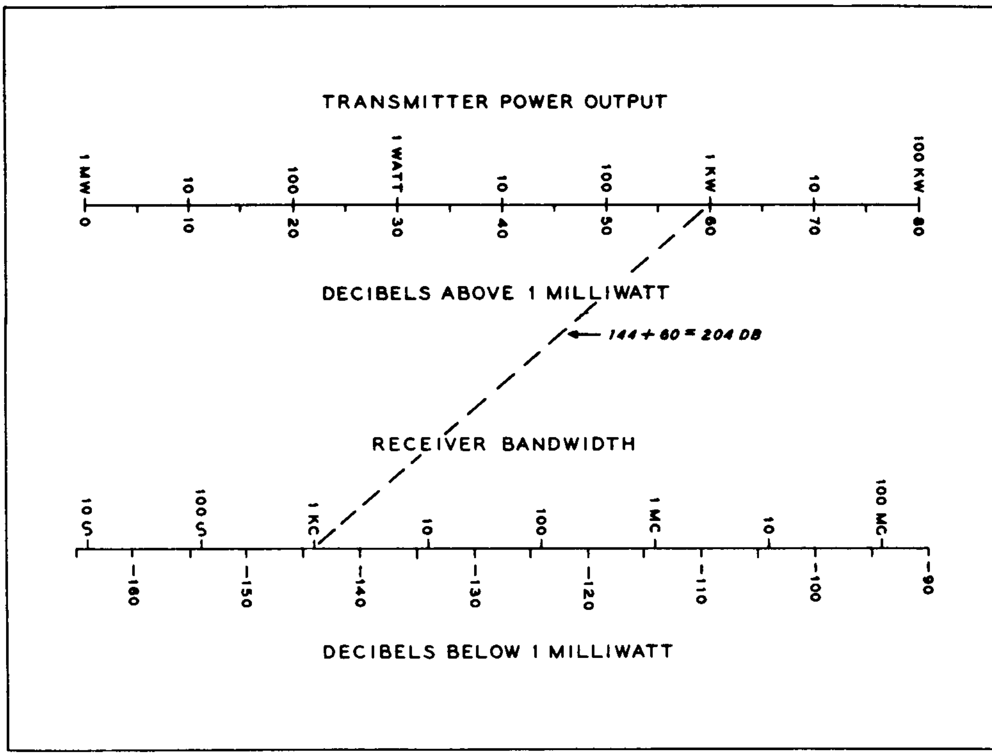


Fig. 6 The "Power Spread" is the amount of circuit attenuation between transmitter and receiver which will result in a S/N ratio of zero decibels.

The Power Spread

If it is possible to determine the value of transmission path attenuation, the *power spread* between the transmitter and receiver may be computed. The power spread may be taken to be that amount of circuit attenuation permissible between the transmitter unit and the detector unit which will result in a S/N ratio of zero db for the path in question. The graph of Figure 6 is useful for this computation. Using a power level of one milliwatt (.001 watt) as a reference, this graph will indicate how many decibels above one milliwatt any given transmission power level will be. The graph also illustrates how receiver bandwidth directly affects the noise output of the receiver. In order to determine the true power spread of any VHF circuit, the noise figure of the circuit must be included. For example: A transmitting system with an effective radiated power of 1000 watts is (from Figure 6) plus 60 db above the reference level of one milliwatt. A receiver with an effective bandwidth of one kilocycle will receive antenna noise having a power level of 144 db below one milliwatt. The power spread is the sum of these two levels, or 204 db. If the receiver has a noise figure of 5 db, the true power spread will be reduced by that amount to a figure of 199 db. This means that if an imaginary attenuator with a transmission loss of 199 db was connected between the transmitter and the receiver, the signal at the receiver would just match the noise level of the receiver. If the attenuator loss was reduced to 190 db, a S/N ratio of 9 db would be observed at the receiver. In an actual VHF communication circuit the transmission path takes the place of this imaginary attenuator.

A factor influencing the transmission path loss is the antenna gain at each end of the circuit. The loss of the path will be reduced by the sum of these gains. For example: If the path loss is equivalent to 220 db, and there is an antenna gain of 20 db (referred to an isotropic radiator) at one end of the path, and an antenna gain of 10 db at the other end, the effective transmission loss will be 220 db minus 30 db, or 190 db. With these assumed conditions, and those set forth in the previous example, a S/N ratio of 9 db will be observed over this communication circuit.

Modulation Systems

The communication circuit is also affected by the amount and type of intelligence transmitted over it. There are several types of modulation systems available for integrating information to be transmitted over a VHF link. The information may be broken down into such forms as: 1—Coded systems employing A1, A2 or F1 emission. 2—Voice modulation, either amplitude or frequency modulated. 3—Video modulation using A4 or A5 emission. The most efficient systems in terms of required bandwidth are those in the first group. Frequency shift keying (F1) is an advanced coding system for the transmission of information, and offers a circuit S/N ratio improvement of as much as 8 db over simple c-w keying. Even more complex systems employing pulse techniques or binary coding have recently been developed, allowing a further increase in the amount of intelligence transmitted in a given bandwidth. For general amateur communication, voice modulation or c-w telegraphy is usually preferred, even though circuit reliability is considerably degraded over other types of signal modulation.

Thus it may be seen that the VHF communication circuit is a complex item composed of many factors which play a vital part in the determination of the degree of maximum circuit efficiency. The VHF engineer must pay exact attention to each factor in the circuit, since a weakness of any one part of the circuit will result in deterioration of the intelligence transmitting ability of the whole circuit. Minute attention to details will pay large dividends in the VHF region.

CHAPTER II

VHF Tropospheric Propagation

As a result of the tremendous increase in VHF activity during the recent years much has been learned about the different modes of wave propagation at these frequencies. The major types of propagation are: 1—Tropospheric Refraction, 2—Tropospheric Scatter, 3—Ionospheric Scatter, 4—Ionospheric Reflection, 5—Aurora Reflection, 6—Meteor Trail Reflection, and 7—Moon Reflection.

Of particular military and commercial interest are the first three of these modes, since they are showing an amazing ability to provide extremely reliable VHF communication circuits over great distances. These circuits require high power and great antenna gain, but provide even more reliable and consistent results than do the lower frequencies, since VHF "scatter" propagation is not seriously affected by the vagaries of ionospheric reflection, magnetic storms, etc. which continually plague the longer wavelengths.

The last four mentioned modes of VHF propagation are generally erratic and unreliable, and are therefore only of academic interest to military and commercial communications. But to radio amateurs who do not require consistent circuits they provide a generous variety of interesting transmission paths. There is still much to be learned about VHF propagation, and radio amateurs are famous for their ability to break down the long "impossible" paths. Undoubtedly the future holds many new, exciting VHF developments, and the "ham" will be in the forefront of such progress.

In the next two chapters the reader will be introduced to the more important aspects of the VHF transmission path. It would be quite possible to devote many more pages to each of the fascinating modes of VHF propagation, but within the space of this text it is necessary to offer the subject matter in condensed, "meaty" form. Much of the information is presented in nomographs and charts which are worthy of careful digestion by the reader.

Nomographs are included at the end of this chapter covering circuit requirements of a tropospheric scatter link, and a summary of various modes of VHF propagation is included at the end of Chapter Three. In addition, a list of references is also included at the end of that chapter.

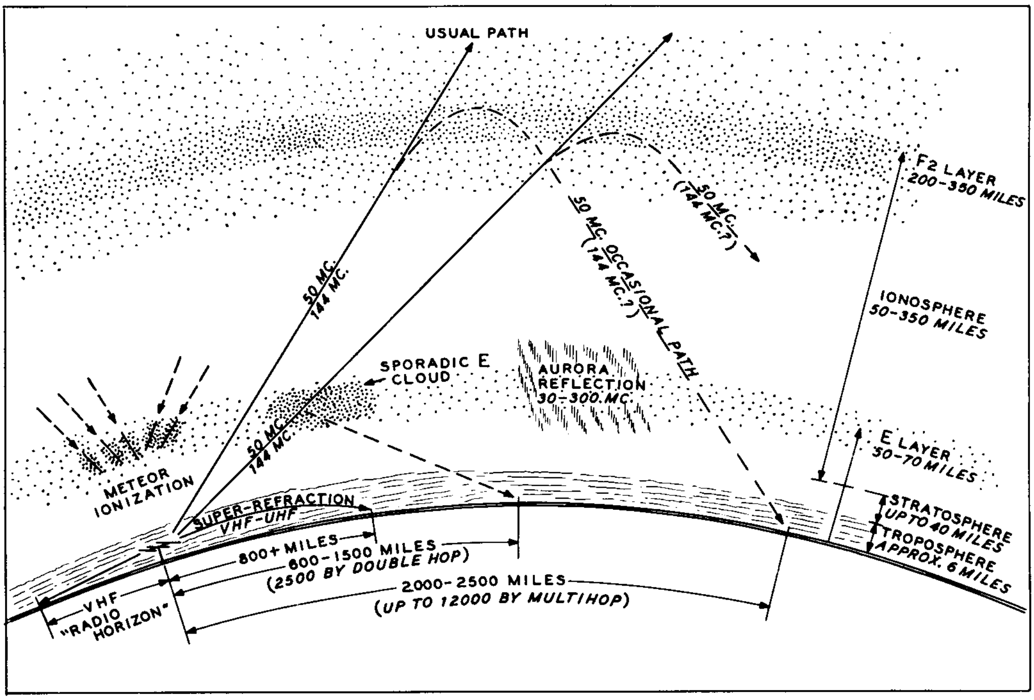


Fig. 1 The atmosphere of the earth is concentrated in a thin layer about three hundred miles thick. Ionized layers of air within this span have the ability to reflect electromagnetic energy of certain frequencies. The atmosphere is divided into strata named the troposphere, the stratosphere, and the ionosphere. It is in this latter region that radio reflection takes place.

The Propagation Medium

All radio waves are propagated through the atmosphere of the earth, which may be considered to reach a height of some three hundred miles. The presence of the atmosphere is not necessary for propagation to take place, as the radio wave travels with equal ease through the vacuum of outer space. It is important to observe the atmosphere, however, as it exerts a profound effect upon the radio wave passing through it. The air around us is indeed a complex substance, composed mainly of oxygen and nitrogen, but also containing traces of helium, neon, xenon, methane, carbon monoxide, ammonia, and nitrous oxide. Supported in the atmosphere and carried from place to place by the varying winds are a host of foreign materials such as dust, ash, pollen, bacteria, water, and fragments from outer space.

The composition of the atmosphere is relatively constant from sea level to the upper reaches, with the density of the air slowly diminishing with increasing altitude. At a height of five or six miles above the earth, man cannot live without an auxiliary supply of air, and at fifty miles the atmospheric density is much rarer than the best vacuum ever produced on earth. This thin blanket supports all life on earth, protects the earth and all living things from deadly emanations from the sun, and provides a medium for long distance radio communication.

For purposes of discussion the atmosphere is divided into various layers, or regions. These are named the *troposphere*, the *stratosphere*, and the *ionosphere*, and are illustrated in Figure 1.

VHF Tropospheric Propagation

The portion of the earth's atmosphere extending from sea level to a height of about six miles is named the troposphere, or "weather layer." This region is the home of the winds, storms, and rains that continually alter and erode the surface of the earth. The temperature of the troposphere decreases about 20 degrees Fahrenheit per mile of increasing altitude, reaching a minimum value near -58 degrees at the upper limit of the region. Meteorological changes in this part of the atmosphere are responsible for many VHF propagation conditions to be discussed in this chapter. Directly above the troposphere is the stratosphere (constant temperature zone), extending to a height of about 40 miles. This latter region is thought to have little or no effect upon VHF propagation.

When VHF signals are transmitted along the surface of the earth, they do not travel in a straight line, but actually bend down slightly beyond the optical horizon due to absorption by the earth's surface of the lower portion of the wavefront. We might say that the wave is "dragging its feet" along the ground. This phenomena causes the *radio horizon* to be extended as though the diameter of the earth were $\frac{1}{3}$ larger than it actually is. *Surface refraction* such as this will exist to some degree on any large sphere, even without atmosphere. Propagation by surface refraction may be considered to be "ground wave" propagation.

Atmospheric Refraction

Formerly it was thought that the "4/3 radius" radio horizon was the useful limit of VHF transmission, but we know today that the troposphere exhibits a profound refractive effect upon VHF waves. A stratification of air density in this region of the atmosphere can produce VHF bending, and this effect is termed *tropospheric refraction.* The curvature of such a radio path is not constant, but tends to be greatest at areas of sharp discontinuity in the atmosphere. Using a simplified analogy, we may say that the denser air at ground level slows the wave front just a little more than does the rarer upper air, imparting a downward curve to the wave travel. The effective radius range of VHF signals is thus extended well beyond the "4/3 radius" radio horizon in many instances (Figure 1).

In a windless, "standard" atmosphere the temperature and water vapor content decrease steadily with altitude. This normal gradient is shown in Figure 2A. However, in the course of average air movements and weather changes the temperature and water vapor gradients may vary considerably from normal. Quite often a warm air mass will move in above a cooler mass, creating a *temperature inversion* such as shown in Figure 2B. This is a change in air density which will cause bending or refraction of the VHF radio path. If the temperature rise exceeds 2.8 degrees Centigrade per 100 feet of altitude, the refractive effect will be just enough to cause the signal to follow the curvature of the earth. This phenomena is known as *super refraction.* It is very unusual for the temperature gradient to reach this figure, and the condition is not common. Instead, it is the sharp drop in water vapor content which accompanies temperature inversions that is responsible for most cases of super refraction. When this decrease exceeds a figure of 0.5 grams per kilogram per 100 feet of altitude, the wave front

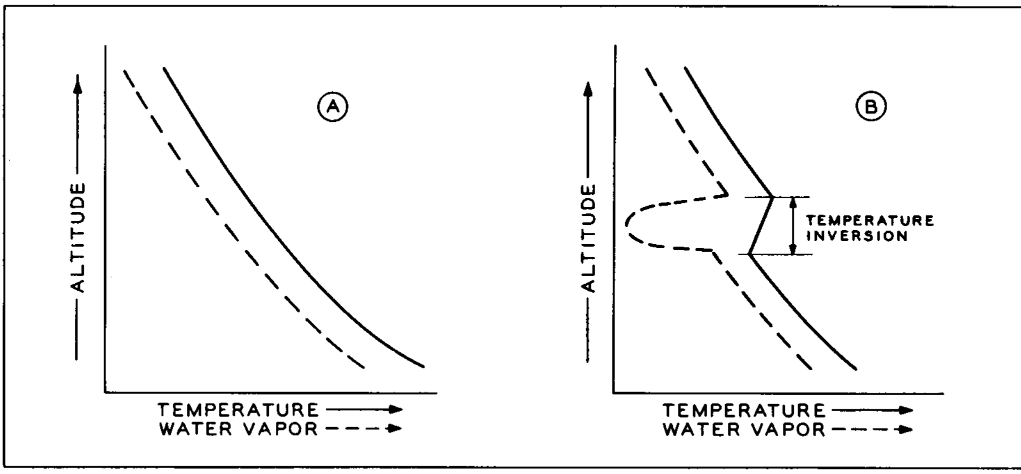


Fig. 2 Temperature inversion is abrupt break in water vapor content of air.

will follow the curvature of the earth and long transmission paths may be covered with very little signal attenuation.

Temperature inversions occur most frequently along coastal areas bordering large bodies of water. This is a result of natural onshore movement of cool, humid air shortly after sunset when the ground air cools more quickly than the upper air layers. The same action may take place in the morning when the rising sun warms the upper air first. These conditions are most likely to take place during the warmer months of the year.

Probably the most spectacular temperature inversions take place at an altitude of several thousand feet. These inversions are created at the junction of large continental air masses which are always moving across the earth's surface. In the United States it is quite common for a relatively cool high

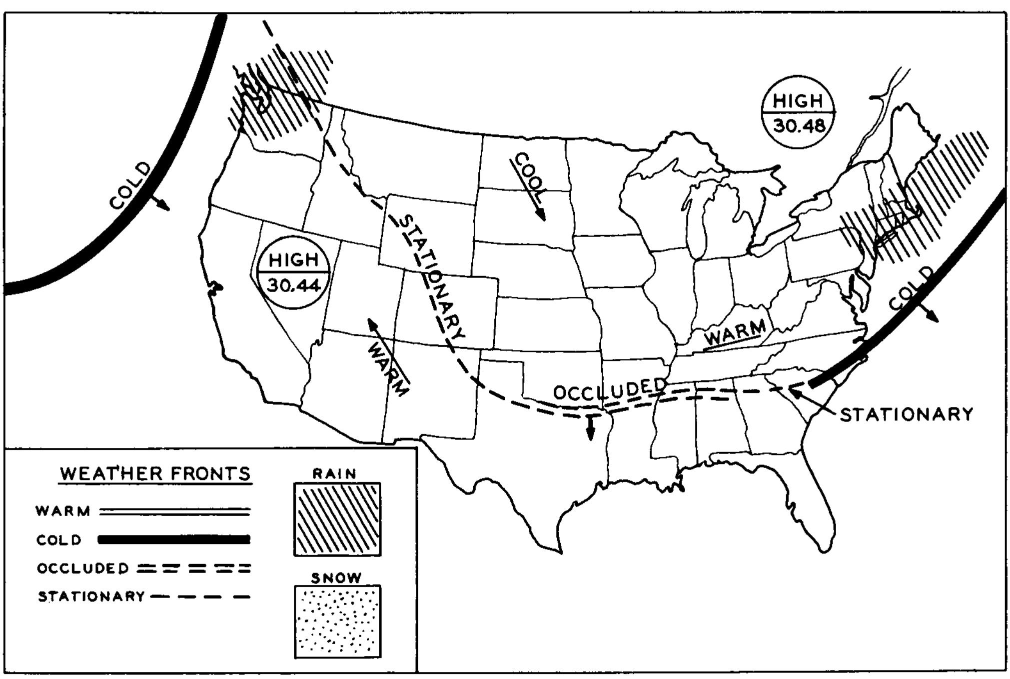


Fig. 3 Temperature inversion forms along stationary air front, as above.

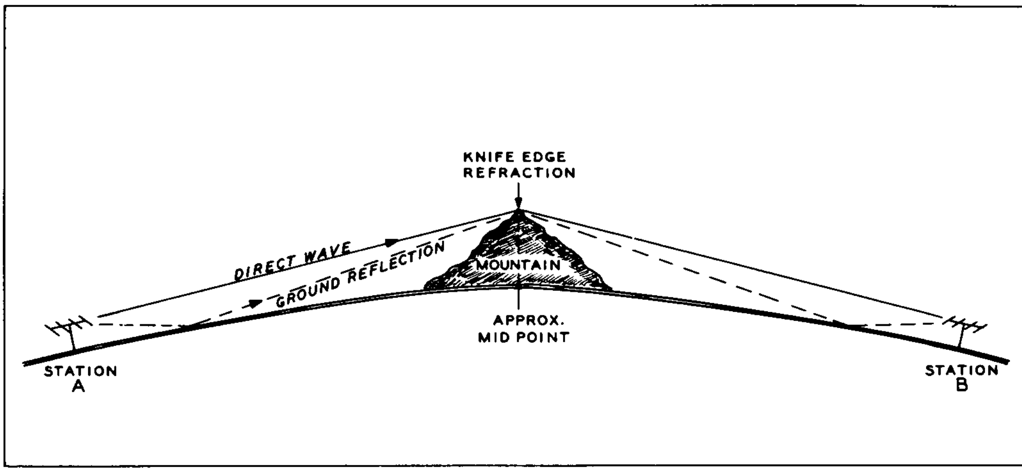


Fig. 4 Obstacle at midpoint of VHF link provides signal boost on long path.

pressure system to move eastward from the Pacific northwest, and for this mass to be met by a warm humid air system moving northward from the Gulf of Mexico. During the summer months and into autumn when these air movements are quite slow, it is possible for a sharp inversion to form along the stationary front at the junction of these air masses, and to extend for as far as 800 miles or more. This condition may last for two to three days, gradually moving eastward and finally dying out as the air systems mix (Figure 3).

Accurate prediction of super refraction is difficult, if not impossible, at least with present weather information. Local inversions can be expected whenever a high barometer reading is accompanied by a rather sudden temperature drop. But the more extensive inversions caused by continental air movements are more complex. The daily weather maps distributed by the U.S. Weather Bureau display surface conditions, and will supply a few clues. However, these maps do not show detailed information on high altitude conditions where the inversions actually take place. The best advice would be to watch VHF propagation carefully, whenever a large high pressure system is moving slowly across the country, and when weather maps indicate formation of a stationary front VHF stations along this front are quite likely to experience super refraction.

Knife-Edge Bending

Under certain conditions it is possible for a ridge of hills or mountains to exhibit very noticeable refraction of a VHF wave traveling over the crest. This phenomena of wave propagation has been demonstrated for years in the Physics Laboratory with light rays, and is familiarly referred to as *knife-edge refraction.* In order to provide an aid to VHF propagation, it is necessary that the hill top runs approximately at right angles to, and midway along the transmission path, as shown in Figure 4. The height of the ridge is quite critical, depending directly upon the distance separating the stations. At a distance of 100 miles a height of about 5000 feet is desirable. The height increases to approximately 12,000 feet for a 200 mile path, etc. Also, antenna height at both transmitter and receiver must be

chosen carefully, as very definite ground reflection patterns will be present. When all factors are optimum, it is possible to realize a large *obstacle gain.* In other words, the received signal will be many decibels stronger than if the mountain were removed, leaving only normal earth curvature. Admittedly, instances of real obstacle gain are not common, but at any rate, hills and mountains need not be considered a barrier to VHF signals.

Forward Scatter

VHF propagation has consistently defied expected limitations. Until a few years ago it was possible to explain occasional long distance phenomena as resulting from super refraction, knife edge bending, or ionospheric skip. But with the advent of higher transmitter power levels, larger antenna systems, and more sensitive receiving equipment, it was found that VHF signals never fall off abruptly below the horizon. Instead, they travel on for surprising distances, and with only a gradual decrease in signal strength. Relatively high transmitter power is necessary to maintain a long distance circuit, especially since the signal fluctuates rapidly over a range of many decibels. The outstanding characteristic of such a "beyond horizon" signal is that the *average* signal level is quite constant (and predictable), regardless of weather or ionospheric conditions. As a matter of fact, it is possible to achieve a more reliable communication circuit with VHF at distances of up to about 1400 miles than is possible with any lower frequency. This new form of propagation is called *forward scatter,* or simply *scatter,* and is separated into two distinct types: One form of scatter takes place in the first few miles above the surface of the earth and is called *tropospheric scatter.* The other type occurs at a height of 36 to 60 miles, in the vicinity of the ionospheric E-layer, and is termed *ionospheric scatter.* A simplified picture of both types of scatter propagation is shown in Figure 5.

Tropospheric Scatter

Tropospheric scatter is caused by random irregularities (or "blobs") in

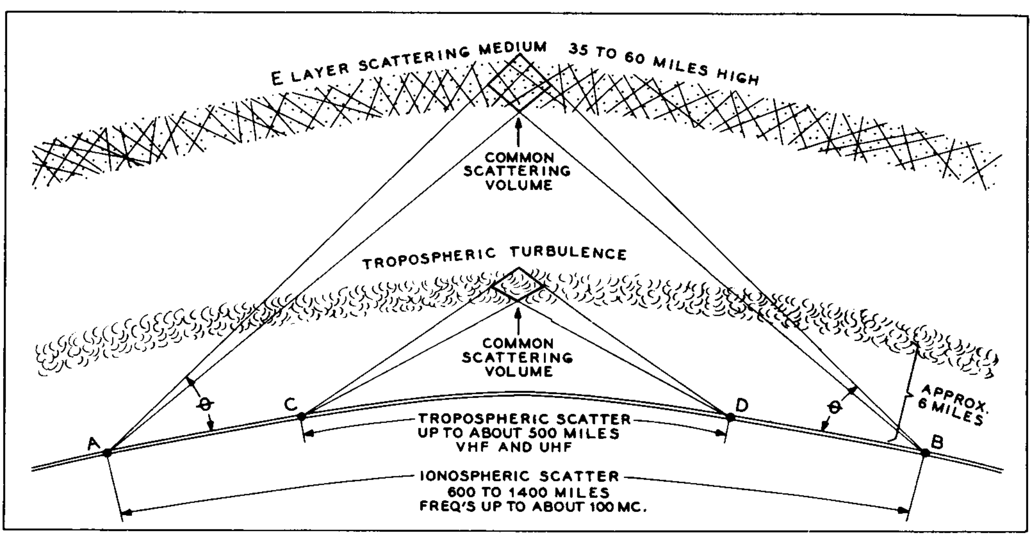


Fig. 5 Common scattering volume in atmosphere links two distant VHF stations.

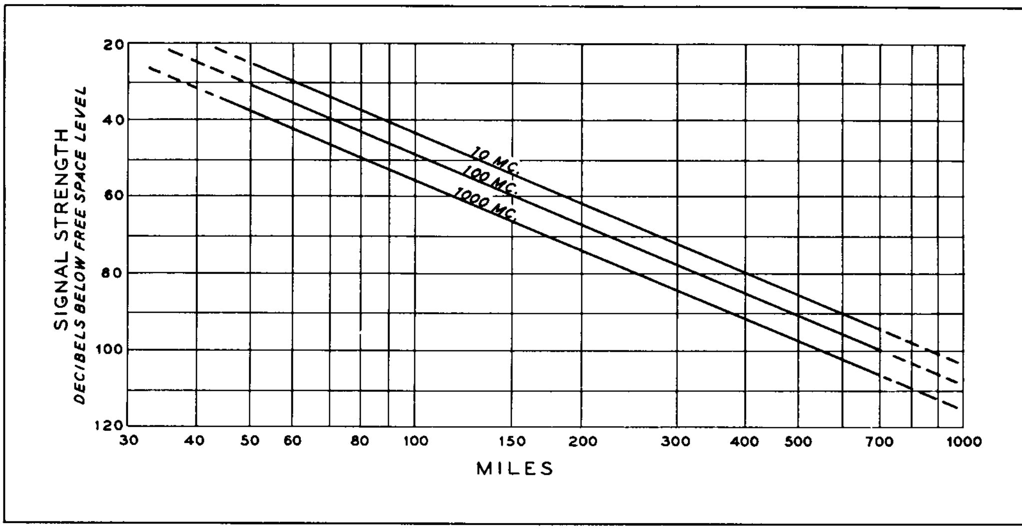


Fig. 6 Tropospheric scattering is possible up to distance of 500 miles. The attenuation increases gradually with operating frequency but rapidly with increase in path length. There is no theoretical limit to signal range.

the atmosphere which are apparently always present. The index of atmospheric refraction is changed sufficiently by these "blobs" to cause faint signal illumination of the ground well beyond the horizon in much the same way that the overhead light beam of a searchlight can be seen from the ground, or the lights of a distant city can be seen as a glow from beyond the horizon. There is theoretically no limit to the possible signal range, although present techniques place the practical limit at about 500 miles for tropospheric scatter. As illustrated in Figure 6, attenuation increases gradually with frequency, but rapidly with an increase in path length. It must be emphasized that the graph presents average values, and that considerable variation may be encountered from hour to hour and with changes in weather, seasons, path topography, etc.

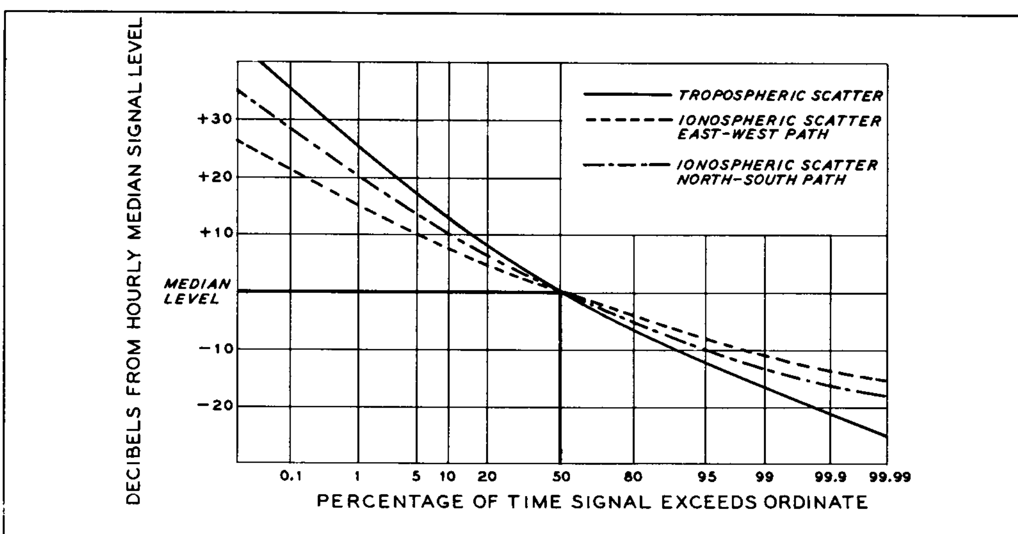


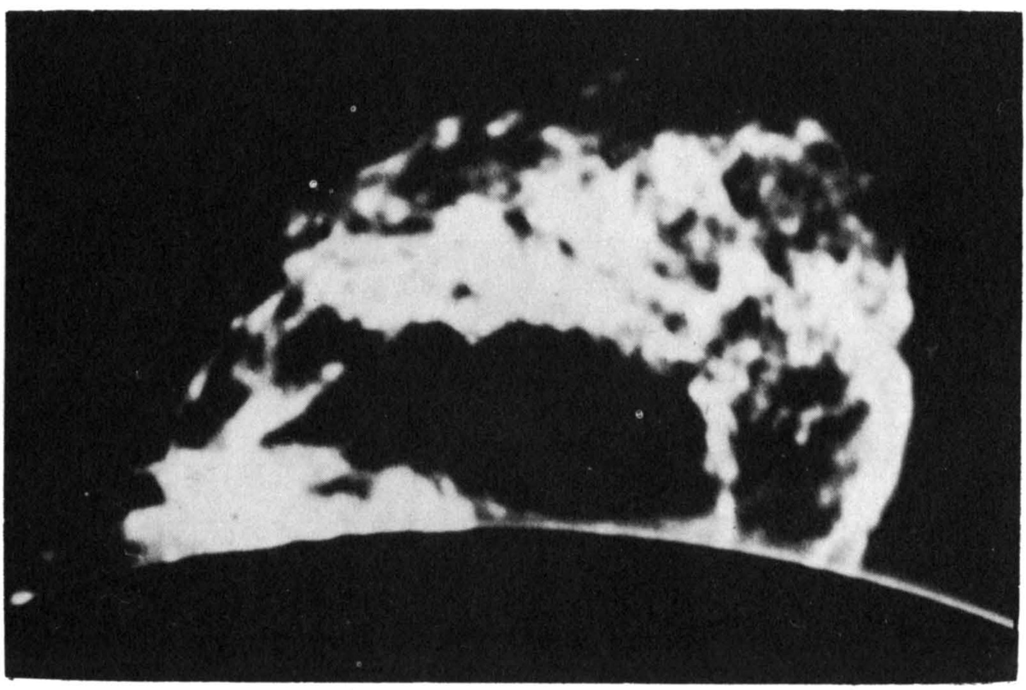
Fig. 7 Ionospheric scattering proves most reliable on an east-west path.

There are two types of signal fading encountered in scatter propagation. The first type is the rapid fade which is caused by multi-path transmission through the atmosphere. The rate of fading increases as either frequency or distance is increased. At 150 mc multi-path fading may drop the signal from maximum strength to minimum, and back to maximum in a matter of seconds. Figure 7 shows graphically the percentage of time the signal will fade to various levels below the average.

The fast multi-path fading will tend to reduce the allowable bandwidth of the communication circuit somewhat, since there may be several signal components arriving at slightly different times at any given segment of the multi path signal. However, indications are that serious difficulty will not be encountered with bandwidths less than 4 mc, as used in TV video circuits.

The second type of fading is the slower type having a period of hours or even days. These slow changes in signal level are the result of variations in atmospheric refraction from day to night, and of humidity and temperature changes along the scatter path.

Figures 8 and 9 are nomographs which may be employed as guides to circuit requirements in tropospheric scatter links. Notice that the examples given with each graph actually fit together to illustrate how a communications system may be designed from the graphs. Notice also how each factor: bandwidth, noise figure, signal-to-noise ratio, transmitter power, antenna gains, and path length affects the other factors. It must, of course, be remembered that the topography of the transmission path may introduce a considerable change in average attenuation. In general, however, the nomographs will prove fairly accurate even over hilly terrain.



Ultraviolet radiations from the sun produce ionization of atmosphere of the earth capable of reflecting electromagnetic radiation. Shown above is tremendous burst of energy in 11 million mile corona discharge on the sun. Increase in radiation in such outburst will cause radio "blackout" on earth.

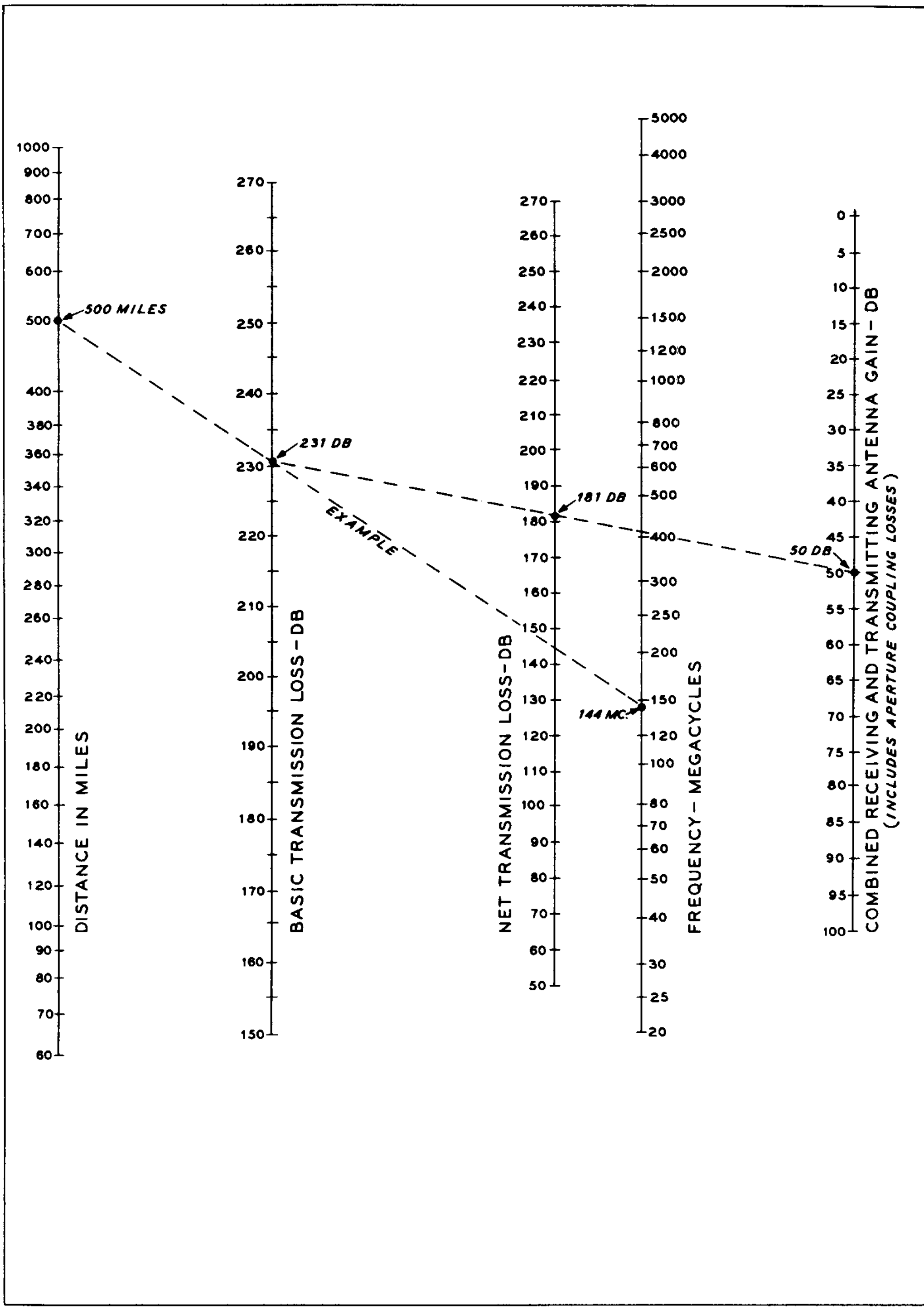


Fig. 8 The nomograph may be used to determine circuit requirements in tropospheric scatter links. As shown, a 144 mc, 500 mile link has a transmission loss of 231 db. If the combined receiving and transmitting antenna gain is 50 db, the net transmission loss is 181 db. This attenuation figure will vary with the topography of the transmission path, but the nomograph will prove fairly accurate even over hilly terrain. The net transmission loss figure obtained from this graph may be used with the nomograph of Figure 9 to obtain the equipment requirements for the path in question.

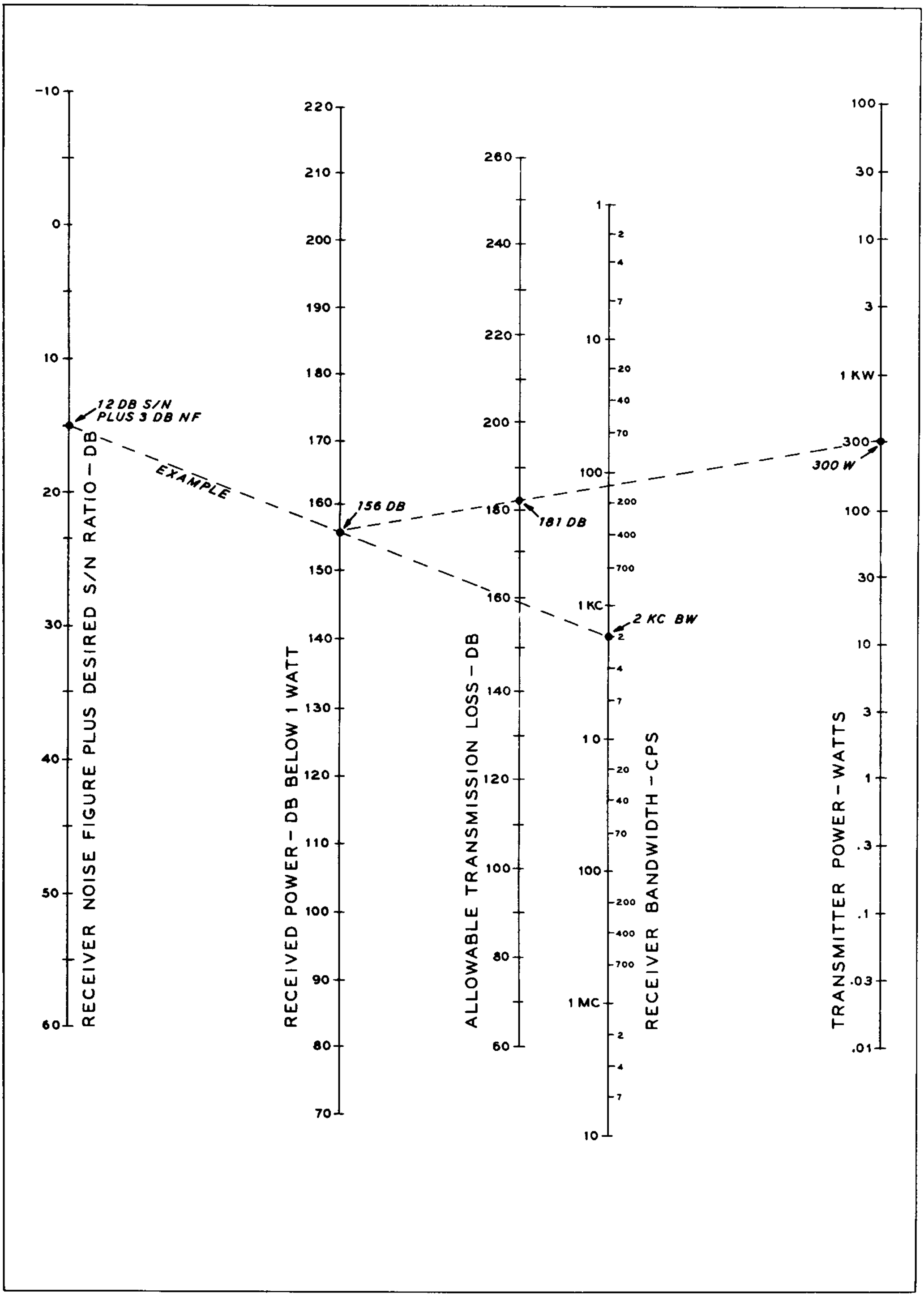


Fig. 9 This nomograph is used to determine equipment requirements for a VHF path when the net transmission loss figure is known. For example, a loss figure of 181 db and a given transmitter power of 300 watts will produce an average signal 156 db below 1 watt at the receiver. If the bandwidth of the receiver is 2 kilocycles, the average level of signal-to-noise for this path will be 15 db. If the noise figure of the receiver is 3 db, the received signal will be 12 db over the noise level. This nomograph should be used in conjunction with that of Figure 8 for complete circuit requirements.

CHAPTER III

VHF Ionospheric Propagation

Most high frequency communication below 30 mc makes use of the ionosphere to reflect radio signals back to earth at distant points. A cross section view of the earth's atmosphere has been shown in Chapter 2, Figure 1, illustrating the various portions of the complex bit of air enveloping the planet. The ionosphere is ordinarily "transparent" to VHF signals, so instead of returning to earth, waves directed at this region continue into the voids of outer space. Only a small fraction of the signal power may be "scattered" back to earth to provide long distance communication. However, there are times when VHF signals are reflected by the ionosphere, and although this may prove to be a nuisance to commercial and military circuits, these occasions provide the amateur with considerable excitement.

The Ionosphere

The ionosphere extends from the lower region of the *mesosphere* to a height of some 350 miles above the earth's surface. Bombardment of this

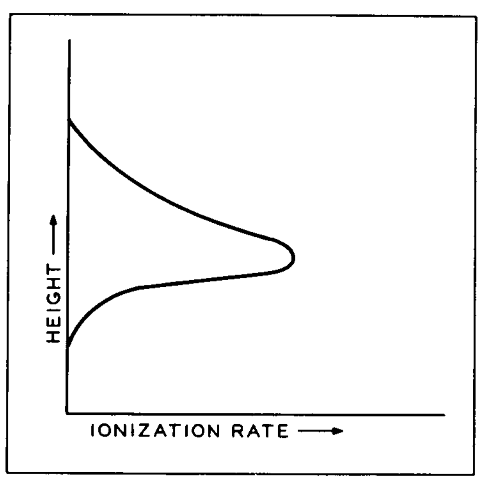


Fig. 1 Free electrons created in the atmosphere by ultra-violet radiation of sun tend to congregate in wide band of maximum density at elevation of 150 to 250 miles above the surface of the earth.

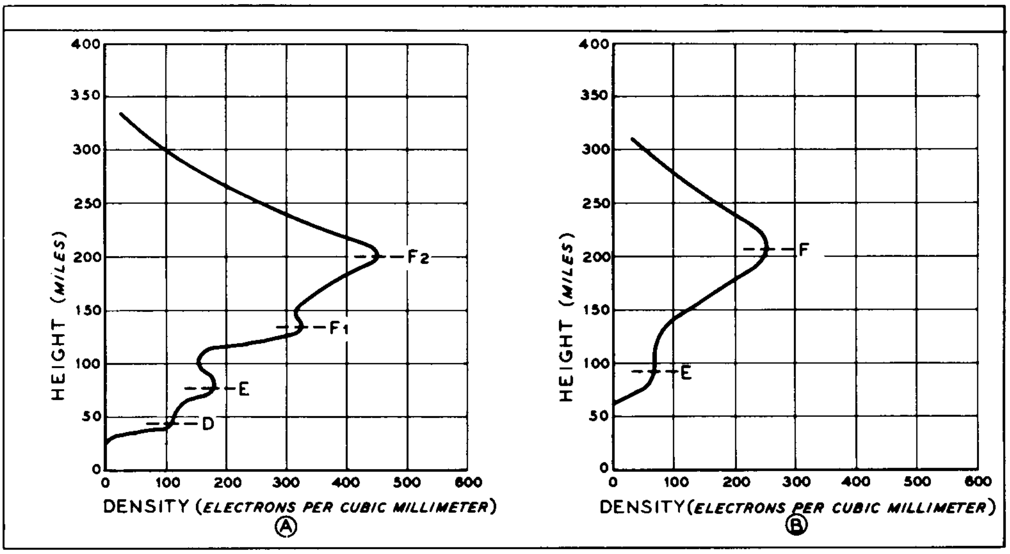


Fig. 2 During daylight hours, four definite electron layers are noted (A). At night, layers draw together leaving only E- and F- layers as shown in (B).

region by ultraviolet radiation from the sun produces ionization of nitrogen and oxygen molecules of the atmosphere. The ionized particles release electrons which, until recombined again with a positive ion, are capable of being set into a state of oscillation by radio frequency energy. The electron will reradiate this energy or perhaps merely dissipate it, depending upon the degree of ionization of the immediate area and the frequency of the radio wave. The rate of creation of free electrons in the ionosphere is shown in Figure 1. A study of the density of free electrons throughout the ionosphere would show that they tend to congregate in bands of maximum density. These bands have no well defined boundaries, blending one into the other. They are called the *D-*, *E-*, and *F-layers* of the ionosphere.

Forever restless and in motion, during the hours of the night when the area is shielded from the radiation of the sun, the layers draw together into a single blanket (F-layer) which hovers at a height of 150 to 250 miles above the earth. At these extreme altitudes the atmosphere is extremely thin and the layer is tenuous and weakly ionized. The recombination rate of ions and electrons is so low, however, that the layer retains a relatively high electron density during the dark hours of minimum energy absorption. The high frequency radio signals easily penetrate the F-layer to be lost in outer space. Lower frequency signals are partially reflected from the F-layer, permitting nighttime communication by ionospheric skip at frequencies below 8 or 10 mc. During periods of maximum sunspot activity the nighttime ionization of this layer is often intense enough to permit ionospheric skip conditions to exist up to 20 or 30 mc, particularly in the warm, tropic areas bordering the equator.

In daylight hours when the ionosphere is exposed to the full force of ultraviolet radiations from the sun the F-layer splits asunder, forming the F_1- and F_2- layers. The lower F_1-layer adds little to the reflection efficiency of the ionosphere, but merely serves as an additional absorber of radio energy that has been reflected from the F_2-layer.

At an altitude of 30 to 40 miles above the earth is another broad, ionized region designated the daytime D-layer. This is an area of relatively dense

atmosphere. The ionization of this layer is directly affected by the quantity of sunlight that falls upon it. The ionization is greatest around noon, and quickly drops to nothing when the sun passes behind the earth. The D-layer is a region of absorption of medium frequency signals, limiting daytime communication on these frequencies to relatively short distances.

THE E-LAYER

Pulsing above the D-layer at a height of 36 to 60 miles is another definable region of ionized atmosphere called the E-layer. The intensity of ionization of this layer follows the sunlight as does that of the D-layer. The nighttime E-layer is porous and sporadic, becoming more intense during the hours of sunlight. The E-layer also absorbs a certain portion of low frequency radiations, but on a smaller basis than the D-layer. The ionization of the E-layer is much greater in equatorial regions of the earth, and will return higher frequencies to earth than is possible in lower latitudes. Shown in Figure 2A is a simplified representation of the electron density of the ionospheric layers during daylight hours, and a typical distribution of the layers during nighttime is shown in Figure 2B.

The locations and idiosyncracies of these various layers were established over the years by scientists working with special radio equipment designed to probe and expose the secrets of the layers. Pulsed signals were projected skywards to the ionosphere and the time interval the signal took to reach a particular layer and to be reflected back to earth was noted. This technique defined the height of the layer and the degree of ionization of each layer. The highest frequency at which a vertically projected signal would return to earth is termed the *critical frequency*. It may be considered to be the *maximum useable frequency* (MUF) for a zero length path between two adjacent earth points.

On any given path, there is a MUF above which communication by ionospheric skip signals was long thought to be impossible. It is now known that the MUF boundary (commonly thought of as the division line between

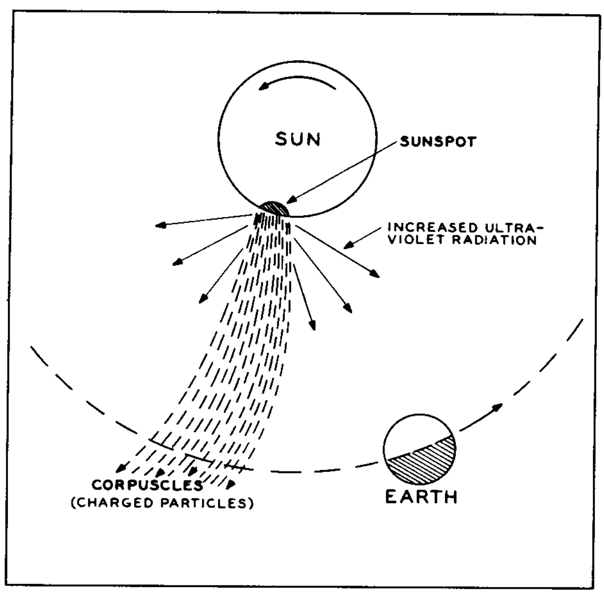


Fig. 3 Sunspot storm bombards earth with charged corpuscles creating ionospheric chaos. The MUF radio dropout is sometimes avoided by earth passing ahead of corpuscular cloud.

"high frequency" and "very high frequency" waves) is not a well-defined demarcation point, but a broad frequency region fluctuating between waves of medium frequency, and those well up into the VHF portion of the spectrum. The MUF is governed largely by solar radiation, and is subject to wide variation from hour to hour, day to night, and season to season. During periods of maximum solar activity the MUF will advance towards 50 mc or so, and will drop well below 10 mc on certain paths during periods of minimal sunspot activity. Thus the lower boundary of the VHF region is blurred and indistinct, as the MUF of ionospheric skip waxes and wanes.

Sudden Ionospheric Disturbances

Besides the more or less regular variations of the ionosphere there are also sudden flare-ups and storms on the face of the sun that apparently are connected with sharp increases in ultraviolet radiation. The storms are also accompanied by the emission of electrified particles or "corpuscles" which bombard the atmosphere of the earth. The increase in ultraviolet radiation is responsible for *sudden ionospheric disturbances* (SID's), while the corpuscular clouds following at a slower speed create delayed but longer lasting ionospheric and magnetic disturbances (Figure 3). During these solar eruptions the MUF will drop quite sharply, even to a near blackout of high frequency reception. The drop is usually preceded by an increase in the MUF which may extend as high as 50 or possibly 100 mc. During this period of high MUF it is possible to receive single- and even multiple-hop F_2 signals in the 50 mc amateur band and in the lower TV channels.

The storms on the sun can often be seen with the naked eye, through a piece of dark glass, appearing as small dark spots on the surface of the sun. These *sun spots* are nearly always present even though not always visible without magnification. Gradually and predictably they build up in numbers and intensity over a regular period of about 11 years. During the peak of this cycle when the solar spots are most numerous, the MUF tends

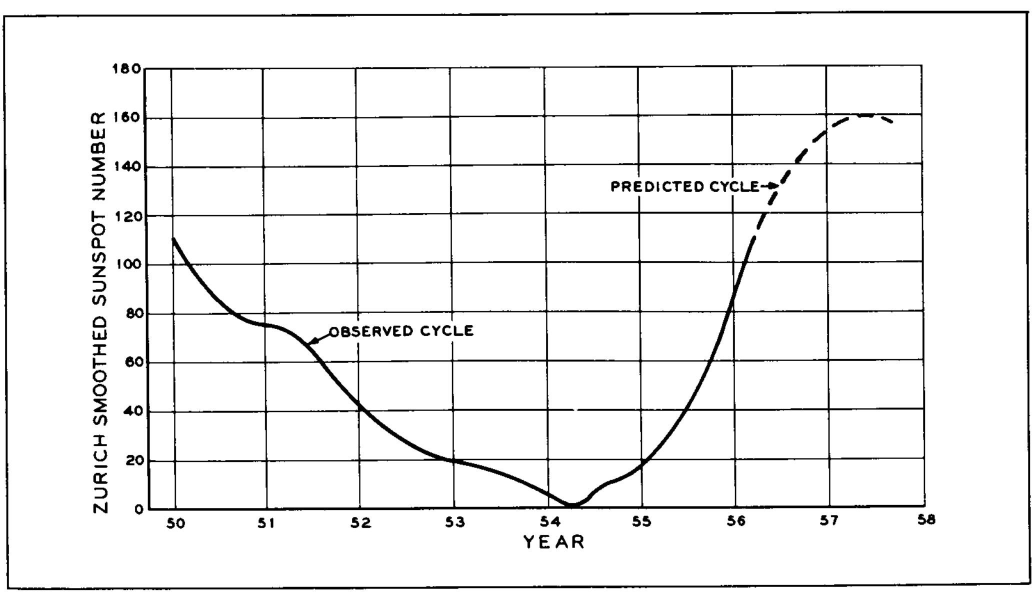


Fig. 4 Sunspot cycle is approaching new maximum at unprecidented rate.

to average quite a bit higher than during periods of minimum sunspot activity. Thus, VHF propagation by *F_2-layer reflection* occurs most frequently during the peak years of the sunspot cycle. The most recent peak occured about 1948, followed by a minimum in mid-1954. The next peak year is expected to be about 1957-58, as shown in Figure 4.

VHF F_2-layer Skip

The corpuscular emissions which bombard our atmosphere during solar storms are attracted towards the poles by the earth's magnetic field and therefore have a most pronounced effect on ionospheric conditions in the higher latitudes. Near the equator an ionospheric storm may hardly be noticed. This helps explain why F_2-layer skip of VHF signals occur most frequently along paths which run roughly north and south and which cross the equator.

Since the MUF reaches its maximum during daylight hours and in the winter months, these are the best times to watch for possible F_2-layer VHF skip signals. However, winter in the northern hemisphere is summertime in the southern hemisphere, so it is necessary to compromise, and be on the lookout during spring and fall months for multi-hop F_2-layer skip between the two hemispheres.

Sporadic-E Layer Propagation

Quite frequently clouds of very high ionization are found in the region of the E-layer. Even though the effect of these clouds is quite well known, their cause is still subject to speculation. There have been attempts to prove a correlation between Aurora and Sporadic-E displays, but the connection is not at all certain. Many times the Sporadic-E clouds achieve a density sufficiently high to reflect VHF signals as high as 70 or 80 mc in frequency. The Sporadic-E skip distance varies from a minimum of 600 miles to a maximum figure of about 1400 miles for single-hop transmission. Some instances of double-hop have been noted. At times the Sporadic-E clouds seem to be quite small and spotty, producing only localized openings, while at other times the ionization covers a large area several hundred miles across (Chapter 2, Figure 1). The clouds have the ability to shift very rapidly and to disappear and reappear in a short span of time. They occur both at night and during daylight hours, but tend to peak about four hours after sunrise, and again just after sunset.

Sporadic-E propagation is possible a high percentage of the time near the equator, and much less frequently near the poles. In the temperate latitudes the spring and early summer months provide peak activity. There is still some question as to how high a frequency may be propagated via Sporadic-E reflection. It is generally thought that 1200-1400 mile reception of TV stations as high as Channel 6 (82-88 mc) is caused by Sporadic-E. However, whether similar distances covered once or twice by amateurs at 144 mc was by Sporadic-E skip or by very high and extensive temperature inversion is not definitely known. Sporadic-E clouds reflect frequencies well below the VHF range, and scatter sounding techniques developed by Oswald Villard, W6QYT have traced definite 14 mc sporadic-E reflections that persisted for periods of many hours.

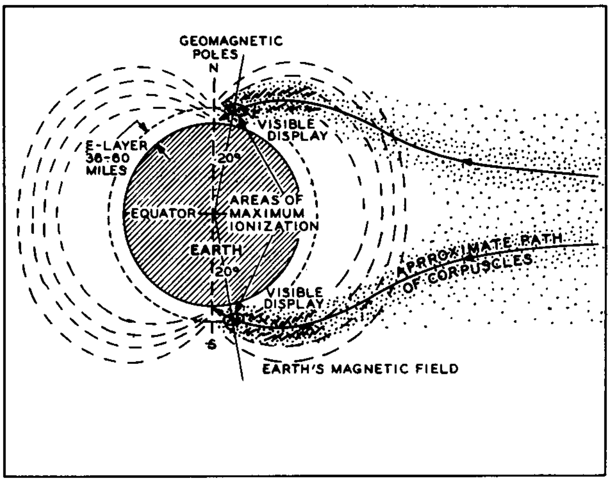


Fig. 5 Maximum aurora display is concentrated in a broad belt near 20 degrees latitude from each geomagnetic pole. Visible ionization occurs at height of 60 to 70 miles in the E-layer.

Aurora Propagation

Another form of VHF propagation which provides excitement for the amateur at the same time it creates a nuisance to commercial circuits is *Aurora reflection.* In an ionized, fluorescent ring crowning the north pole is the Aurora Borealis, and a similar ring circling the south pole is called the Aurora Australis.

The generally accepted theory covering these awesome displays is that the ionization is caused by clouds of charged particles thrown off by the sun during solar storms. These particles are the same ones thought to cause ionospheric and magnetic storms. These "corpuscles" of energy are captured by the earth's magnetic field and follow the magnetic lines of force toward the geomagnetic poles, ionizing the upper atmosphere as they collide with its atoms. Most of this visible ionization occurs in the E-layer at a height of 60 to 70 miles, and at about 20 degrees of latitude from each geomagnetic pole. But many of the colorful rays and streamers flash at heights up to 700 miles and can sometimes be seen overhead as far south as the 40th latitude in the eastern United States, and the 45th latitude in the western part of the country. The display is visible almost constantly in the Aurora belts circling the poles (Figure 5).

The more heavily ionized portions of the Aurora will reflect VHF signals, and communication between two distant points may be established by pointing the transmitting and receiving antennas northward toward the Aurora belt. The Aurora-reflected signals will be severely modulated or distorted due to the rapid "oscillation" of the Aurora, resulting in a characteristic "growl" or "hiss" superimposed upon the signal. Voice modulation becomes very rough and practically unreadable at 50 mc, while at 144 mc modulation is almost completely wiped out by "Aurora Hiss." It is therefore necessary to employ c-w telegraphy at slow speeds for Aurora communication above 144 mc.

Even though the visible portions of an Aurora display may extend above the northern horizon to a considerable height (at times reaching overhead and into the southern sky) it seems that VHF reflection takes place strictly from E-layer ionization in an area lying several hundred miles north of the observer and close to his horizon. Experiments with tilted antennas have yielded no significant deflection from the streamers and curtains which

create the colorful spectacle visible at the higher angles of elevation.

Since Aurora is caused by emission of charged particles from the sun during solar storms, it is natural to find that Aurora-skip effects follow the sunspot cycle and reach a peak at the same time. Auroral propagation decreased as the sunspot minimum was passed in 1954, and will be on the increase until the next maximum period of sunspot activity, estimated to be during 1957-58. Maximum Aurora activity occurs in the spring and fall months, especially March and September, although there may be openings during any month. Aurora signals will reach maximum strength at about sundown, and again around 2 to 3 a.m. The farther north a station is situated the more frequently it will encounter Aurora propagation. Figure 6 indicates the estimated number of days per year when Aurora openings will occur for various parts of the United States and Canada at the peak period of the sunspot cycle. During rare occasions it may be possible to employ Aurora propagation in the southernmost portions of the United States, but ordinarily such conditions are limited to the northern states.

Because of the shallow north-south dimension of the Aurora belt, east-west transmission paths are usually favored. At times it is possible for stations up to 800 miles apart to communicate via Aurora skip. Generally speaking, the eastern station should aim his antenna slightly west of magnetic north, while the western station aims slightly east of magnetic north. However, if the antenna has a fairly sharp directional pattern (say, 25 degrees or less) it is wise to search across the northerly direction for strongest signals as there may be some variation and movement of the signal path as the Aurora progresses.

Prediction of Aurora Skip

As previously mentioned, Aurora display is associated with a solar storm. The effects of this storm are usually noticed first on earth as a SID (Sudden Ionospheric Disturbance). About 26 hours after the start of the SID the corpuscular emissions reach the earth and, among other effects, produce

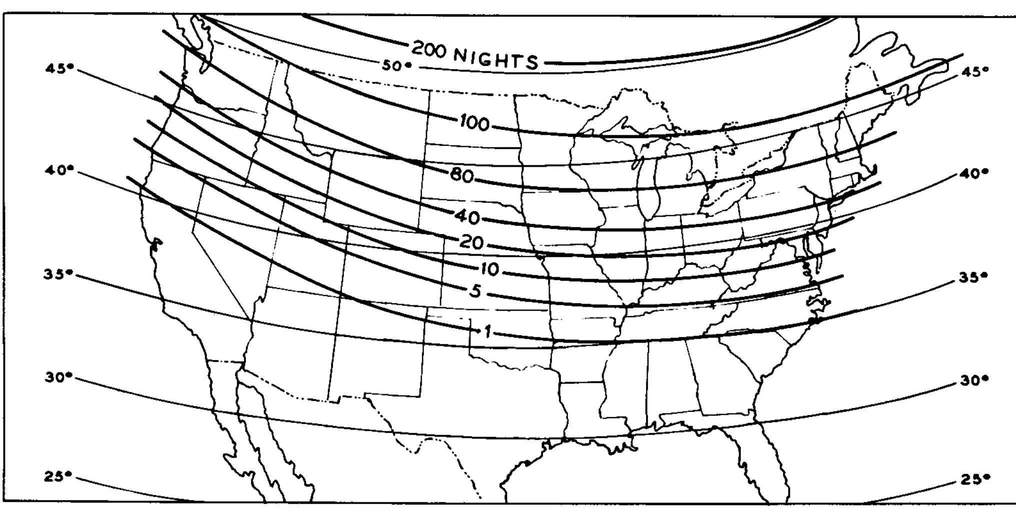


Fig. 6 Aurora display is most prevalent at northern latitudes, but can be seen on occasion as far south as New Mexico or Arkansas, as shown above.

the Aurora. Therefore, we can often predict Aurora as well as magnetic and ionospheric storms 26 hours in advance by watching for SID's. Sometimes the corpuscles miss the earth entirely, so a SID is not always followed by a magnetic storm, but the magnetic storm is invariably preceeded by a SID.

The Bureau of Standards radio station WWV transmits ionospheric storm information daily, once each hour. A series of N's in Morse Code indicate normal conditions, while a series of U's indicate unstable conditions, and W's are warnings of possible ionospheric and magnetic disturbances. It is always wise to watch for Aurora reflection when WWV is sending "W's", or even "U's".

Ionospheric Scatter

Over forty years ago Kennelly (Proc. IRE, July, 1913) considered the possibility of scattering of radio signals in the lower ionosphere. In 1929 Eckersley began investigating what he theorized to be scattering in the region of the E-layer. But it was not until more recent years that serious and concentrated experimenting with *ionospheric scatter* has brought its really useful characteristics to the forefront.

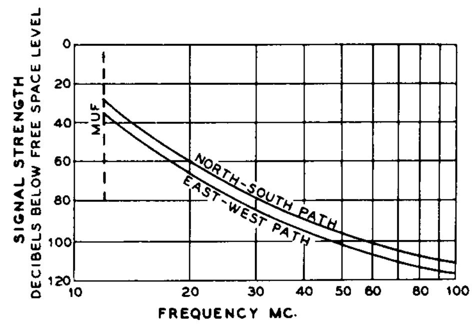
Starting in January, 1950, test transmissions were sent from Cedar Rapids, Iowa to Sterling, Va. on 49.8 mc in a joint project of Collins Radio Co., and the National Bureau of Standards. NBS subsequently enlarged the test program to include longer paths and various frequencies, and has now compiled volumes of important data on ionospheric scatter propagation.

Professional opinions differ considerably as to the mechanism which produces the constant, turbulent, but very weak ionization of the E-layer. One theory suggests that atmospheric turbulence similar to that which causes tropospheric scatter is responsible, while another school of thought proposes that a continuous meteoric bombardment of the atmosphere creates many random ionizations.

Ionospheric scatter propagation via the E-layer is effective only over a limited portion of the frequency spectrum, as contrasted to tropospheric scatter which is not nearly so critical as to frequency. Figure 8 shows the expected circuit attenuation versus scatter frequency for this mode of propagation, and it can be seen that 100 mc is about the useful upper limit. Even above 50 mc transmitter power requirements are rather excessive. The

Fig. 7 Aurora display signifies east-west transmission paths open on 50 and 144 mc up to 800 miles. (Photo taken at Anderson, Indiana.)

Fig. 8 E-layer scatter propagation is effective below 100 mc, showing best properties on the north-south path.



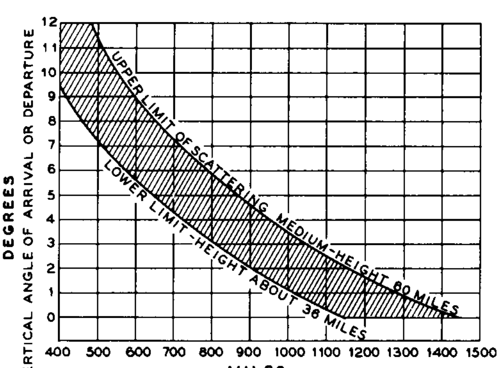
lower frequency limit of ionospheric scatter (about 25 mc) is determined by the masking action of normal ionospheric skip signals. Regular skywave propagation will create undesirable interference and produce selective fading on a scatter link circuit. Paradoxically, we might note that MUF could be taken to mean the "Minimum Useable Frequency" for a scatter link circuit.

E-layer scattering occurs at a height of 36 to 60 miles, near the lower levels of the layer, and produces an optimum skip distance of 600 to 1200 miles. At distances less than 600 miles the angle of incidence of the reflected wave becomes excessive and transmission loss of the circuit increases rapidly. With normal antenna heights and low, grazing radiation angles (Figure 9) the wave will skip out to about 1200 miles. This range can be extended to some 1400 miles by employing antenna sites located several thousand feet high, such as a mountain top, combined with a sea level horizon.

Theoretically, it would be possible to communicate via double-hop ionospheric scatter, which could extend the range to 2000 miles or so, but circuit attenuation cannot be overcome even with the most powerful transmitters and sensitive receivers in use today.

Signals propagated by ionospheric scatter are subject to a rapid fade similar to that noted in cases of tropospheric scatter. The graph of Figure 10

Fig. 9 E-layer scatter range is some 1400 mi. when high antenna and sea level horizon are combined. The scatter occurs at layer height of 36 to 60 miles.



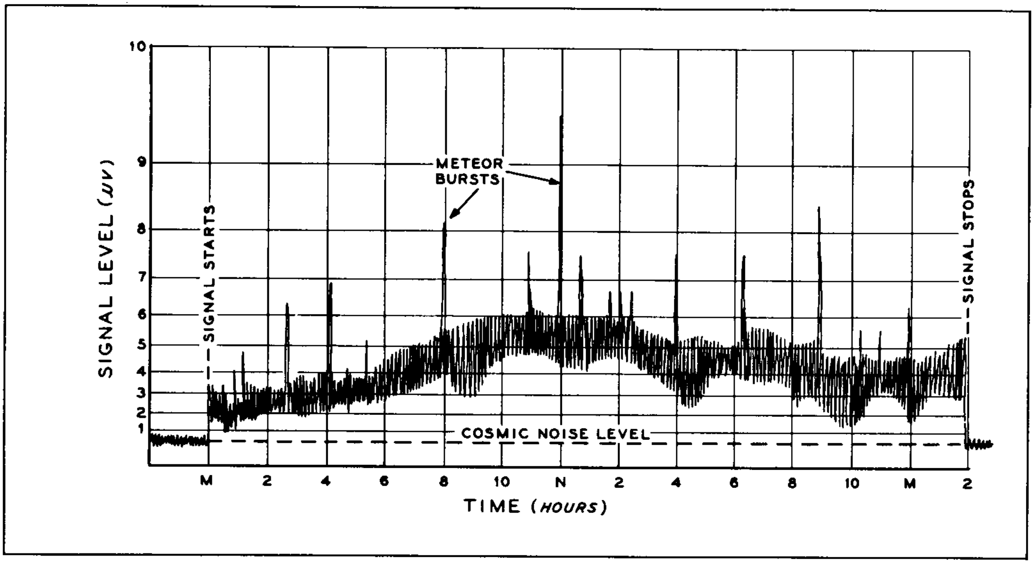


Fig. 10 Scatter signal level is low, punctuated by meteor bursts (above).

shows the depth and frequency of these fades, and illustrates that a "floor level" exists below which the signal practically never fades. For commercial circuits it is this level which must be used as a design center, whereas amateurs could be satisfied with less reliability. A number of other, slower variations of fade are also encountered on scatter circuits. There is a definite tendency for signals to start building up about 6 A.M., and to peak about noon. This effect is most noticable on the shorter paths. It is apparently caused by the combination of maximum ultraviolet ionization of the atmosphere which occurs at noon, and meteor bombardment which peaks about 6 A.M. Signals will also tend to increase somewhat during SID's, Aurora displays, and particularly during periods of Sporadic-E reflection. Summer months yield signals averaging a few decibels above winter signals. Transmission paths at high latitudes are generally stronger than are the paths of middle or lower latitudes, and north-south circuits are usually better than east-west paths.

Antenna Gain on Scatter Paths

When large antennas are employed on long scatter circuits, the effective power gain will be somewhat less than the free space gain of the antenna. This is due to multipath propagation of the signal which causes phase differences in the wave front arriving at various portions of the antenna. This effect is sometimes referred to as *Aperture Coupling Loss,* and the nomograph of Figure 11 illustrates what can be expected.

It is also interesting to note that the received scatter signals do not fade to a lower level on a small antenna as compared to a larger one, and that the only advantage of the large antenna is that the signal peaks will reach a higher value. This effect is apparently caused by the fact that when signals fade deeply they may be arriving from an indirect angle to which the larger antenna is not sensitive, in which case the smaller antenna will be just as effective as the bigger one, as shown in Figure 12.

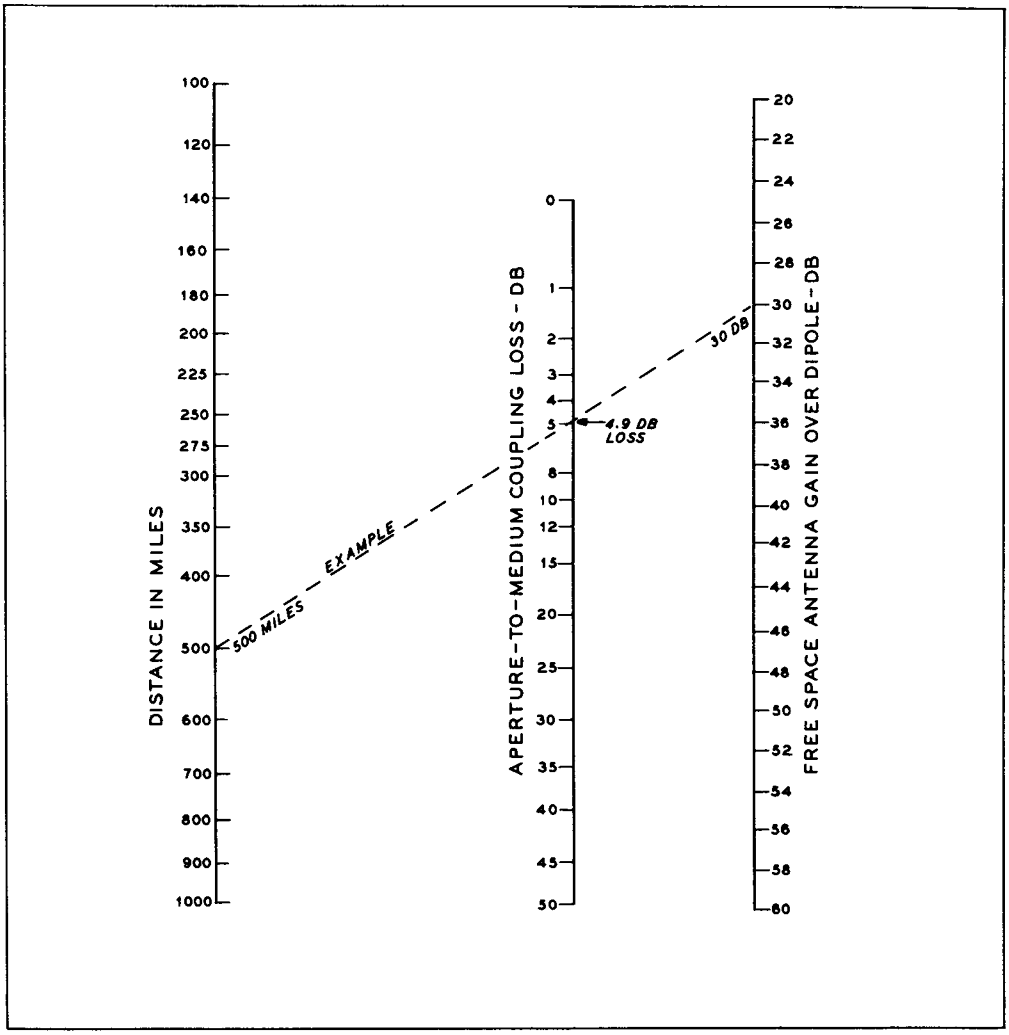


Fig. 11 Aperture coupling loss due to multipath propagation on scatter path.

Diversity Reception

Because of the effects of multi-path transmission, it has been found that signal levels can be held to a higher average value by using two or more spaced antennas, with automatic means for selecting the one which is providing the strongest signal at the moment. A minimum spacing of 15 to 20 wavelengths between the antennas is recommended when the antennas are in a line perpendicular to the transmission path. The chart of Figure 13 shows how much improvement in average signal level can be realized with dual diversity reception.

Meteor Trail Reflection

Not too many years ago it was commonly thought that space was almost a true vacuum, with practically nothing existing in the void between the planets and stars. Meteors have been observed for centuries, but even they

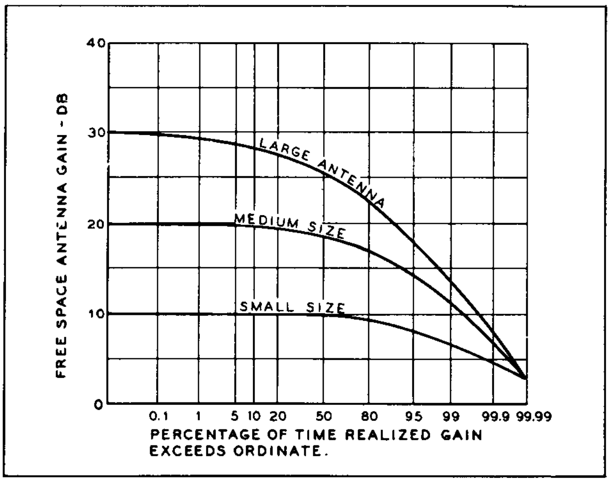


Fig. 12 Scatter signals do not fade as markedly when small antennas are used as compared to fading noticed when high gain arrays are employed on same circuit.

were assumed to be relatively few in numbers, appearing only as a streak of light in the sky when their tremendous speed caused them to burn up in our atmosphere. Today, however, we know that our solar system contains innumerable little particles with which the earth is constantly colliding as it sweeps through the heavens on its annual journey around the sun. Most of these *meteorites* are so small that they are invisible to the naked eye as they burn and evaporate in the ionosphere. Only a very few are large enough to be noticed, and an extremely small percentage are big enough to actually reach the ground before they are completely consumed.

There are two distinct groups of meteorites. The most common type is of a sporadic nature, arriving from random directions and having wide ranges of velocities. The chart of Figure 14 lists these meteorites according to size, mass, approximate number per day, etc. The second type is that encountered in meteor showers which occur when the earth passes through a meteor stream. There is good reason to believe that these streams have a direct relation to comets, and that they are literally the debris which the comet leaves in its orbit around the sun. Meteorites in these streams are moving together at the same speed and in the same orbit. The width of the streams varies considerably, requiring anywhere from a few hours to several days for the earth to traverse. Figure 15 illustrates a typical shower.

DEPTH OF FADE BELOW MEDIAN DB	PERCENTAGE TIME WITH NO DIVERSITY	PERCENTAGE TIME WITH DUAL DIVERSITY
2.9 DB	30 %	9 %
4.9	20	4
8.2	10	1
11.3	5	0.25
15.4	2	0.04
18.4	1	0.01

Fig. 13 Diversity reception overcomes multi-path fading of scatter propagation. An antenna separation of 15 or more wavelengths is required.

Fig. 14 Largest portion of meteors reaching earth are less than 0.016 cm. diameter. Over 12 billion meteors are swept up by earth each 24 hours.

SPORADIC METEORS	VISUAL MAGNITUDE	WEIGHT, GRAMS	DIAM., CM	ESTIMATED NUMBER SWEPT UP BY EARTH EACH DAY
PARTICLES WHICH REACH GROUND BEFORE BURNING UP	-12.5	10,000	16	10
PARTICLES WHICH BURN COMPLETELY IN UPPER ATMOSPHERE	-10	1000	8	100
	-7.5	100	4	1000
	-5	10	1.6	10,000
	-2.5	1	.8	100,000
	0	.1	.4	1 MILLION
	2.5	.01	.16	10 MILLION
	5	.001	.08	100 MILLION
	7.5	.0001	.04	1 BILLION
SMALLEST PARTICLES DETECTABLE BY RADAR →	10	.00001	.016	10 BILLION

Meteor Showers

Until quite recently meteor showers were known to exist only during the night, for the simple reason that it was not possible to observe a meteor against the daylight sky. But, starting in 1947, various investigators began to observe daytime showers by employing radar techniques, finding their hourly rates and occurence surprisingly high. The chart of Figure 16 illustrates meteoric distribution throughout the year 1949-50 in the northern hemisphere. The chart is divided into daytime and nighttime observations, and shows the extensive daytime increase during the months of May, June and July. Although investigation of daytime meteor showers is relatively new, it appears that the summer increase is a recurrent phenomena. It is curious to note that ionospheric scatter circuits enjoy an increase in average signal levels during the same period of the year.

While a shower lasts the number of visible meteors increases from a normal count of two to four per hour to as many as 50 or more per hour, depending on which of the regular showers is occuring. Some nighttime showers in the past have produced tremendous spectacles, with estimated counts running up to thousands per minute for one observer. This number

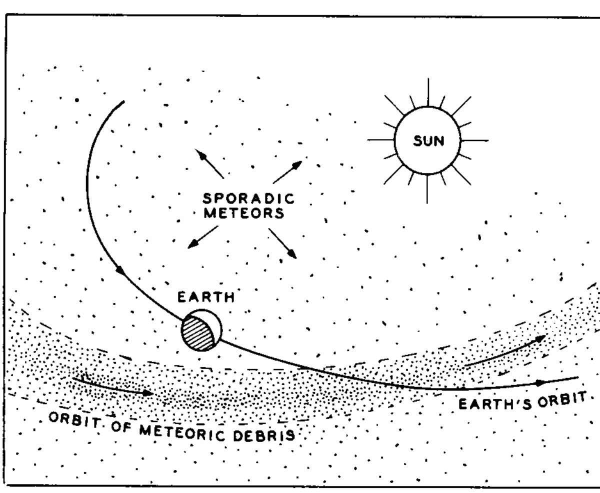


Fig. 15 Earth cuts through a stream of meteor debris in path of comet. Width of debris orbit may require several days for earth to move out of path.

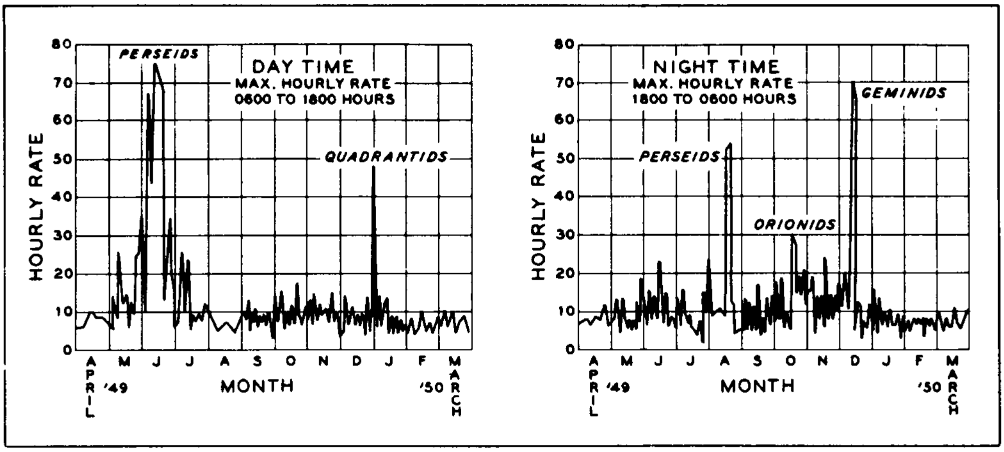


Fig. 16 Hourly rate of large meteor shower may exceed 200 in spring months.

would provide continuous enough ionization to reflect VHF signals over long distances with little attenuation, but unfortunately, the streams causing these huge showers have deviated from the earth as a result of the gravitational effects of Saturn and Jupiter, and at the present time the earth does not encounter any streams of this density. The chart of Figure 17 lists the major meteor showers, dates, average hourly rates, etc.

Because commercial and military ionospheric scatter circuits must be designed to operate during minimum signal periods, factors which tend to increase signal levels for short periods are not of great value, and may often be quite annoying. But to the radio amateur who is ready to leap at any chance to utilize unusual propagation effects, meteor showers can prove quite interesting. Figure 17 lists specific dates for the various meteor displays, but meteoric activity tends to build up and drop off gradually so it is wise to conduct tests a few days either side of the peak. The nighttime displays reach their peaks of intensity sometime between midnight and 6 A.M., while the daytime showers peak between 6 A.M. and noon.

Meteor Ionization

Meteor ionization takes place in the vicinity of the E-layer at a height of about 60 miles. The ionization is produced in a cylindrical column which expands rapidly by diffusion in the rarified atmosphere. It takes only a tiny particle to create sufficient ionization for reflection of VHF radio waves. It is believed that a meteorite with a diameter of only .006 inch is large enough to cause this effect, and it is estimated that the earth sweeps up about 10 billion of this size or larger each day.

The most important type of meteor trail reflection is the *specular,* which is obtained when the trail is perpendicular to the transmission path of the radio signal. This type is the only one which will provide useful reflection from the smaller, *under-dense* meteors. When the column becomes ionized sufficiently to act much in the manner of a metal reflector, it is termed *over-dense,* and will reflect radio energy from almost any angle. Over-dense columns are caused by meteorites about .06 inch diameter or larger, and are less frequent than the under-dense columns, in fact, not frequent enough to support a continuous communication circuit.

	NAME OF SHOWER	DATE OF MAXIMUM	RADIANT COORDINATES		AVERAGE HOURLY RATE *	VELOCITY KM/SEC.
			RIGHT ASCENSION	DECLINATION		
NIGHT TIME	QUADRANTIDS[1]	JAN. 3	230°	+52°	35	39
	LYRIDS[1]	APRIL 21	270	+33	8	51
	η-AQUARIDS[1]	MAY 6	338	+3	12	66
DAY TIME SHOWERS	PISCIDS[2]	MAY 7-13	26	+25	30	—
	O-CETIDS[2]	MAY 21	30	−3	20	—
	ζ-PERSEIDS[1]	JUNE 3	61.5	+24.4	40	28.8
	ARIETIDS[1]	JUNE 8	44.3	+22.6	60	37.6
	54-PERSEIDS[2]	JUNE 25	68	+33	50	—
	β-TAURIDS[1]	JULY 2	86.2	+18.7	30	31.5
	α-ORIONIDS[2]	JULY 12	87	+11	50	—
	ν-GEMINIDS[2]	JULY 12	98	+21	60	—
	λ-GEMINIDS[2]	JULY 12	111	+15	32	—
	β-AURIGIDS[2]	JULY 25	87	+38	20	—
NIGHT TIME	δ-AQUARIDS[1]	JULY 28	339	−11	10	50
	PERSEIDS[1]	AUG. 10-14	47	+58	50	61
	GIACOBINIDS[3]	OCT. 9	262	+54	VARIES	20
	ORIONIDS[1]	OCT. 20-23	96	+15	15	68
	TAURIDS[1]	NOV. 3-10	55	+15	10	27
	BIELIDS[3]	NOV. 14	25	+45	VARIES	22
	LEONIDS[1]	NOV. 16-17	152	+22	12	72
	GEMINIDS[1]	DEC. 13-14	113	+32	60	35
	URSIDS[1]	DEC. 22	207	+77	13	38

1 – *ANNUAL RECURRING SHOWERS.*

2 – *IT IS NOT YET KNOWN IF THESE SHOWERS APPEAR REGULARLY.*

3 – *THESE SHOWERS HAVE PRODUCED TREMENDOUS DISPLAYS IN THE PAST, ALTHOUGH IN RECENT YEARS THEIR HOURLY RATE HAS BEEN QUITE SMALL.*

* – *HOURLY RATE OF VISIBLE TRAILS FROM ONE OBSERVATION POST.*

Fig. 17 Specific dates for the largest meteor showers are listed above. The spring showers peak between midnight and 6 A.M., and again near noon.

Reflections from under-dense columns are of very short duration, and it is therefore necessary to depend upon large numbers of small meteorites to provide an overlapping pattern of reflection. Signals propagated by this mode require high transmitter power and high gain antennas to maintain the circuit through periods of minimum meteor activity. It is this form of propagation which some investigators believe is largely responsible for "ionospheric" scatter.

Probably the first manifestation of meteor reflection noticed by amateurs was the "whistler" effect. This occurs when the signal is reflected from a meteor head which is moving toward or away from the transmitter. Since the velocity of the meteor is quite high, a Doppler shift of several hundred cycles may be noticed on the reflected VHF signal. This change in frequency of the received signal will not be as great if the meteor is not moving directly toward or away from the transmitter, and practically no frequency shift will take place when the meteor trail is perpendicular to the transmission path. Frequency change is greatest when movement is oblique to receiver.

Sporadic Meteors

At about 6 A.M., local time, it may be said that we are "looking straight ahead" in the direction which the earth moves through its orbit. Straight ahead in this case will be (if we are standing at the equator) directly overhead. It is at this time of the morning that our position on the earth will collide with the maximum number of sporadic meteors for the day. At about 6 P.M., or twelve hours later, we will be at a point where meteor velocity must be great enough to "catch up" with the flight of the earth, and meteor collisions will be at a minimum. As a result, meteor trail reflected signals peak at approximately 6 A.M., and drop to a level about 10 to 15 db lower at 6 P.M.

Although sporadic meteors collide with our atmosphere at all angles, the ionized columns tend to take east-west directions because of the earth's direction of travel through space. Since specular reflection is the primary cause of continuous meteor propagation the north-south transmission paths usually suffer less signal attenuation from the east-west columns of ionized atmosphere. It should be understood that this applies only to reflections from under-dense meteors. East-west paths should be as good as north-south paths on reflection from the larger, over-dense meteors.

The power levels and antenna sizes needed for specular reflection are somewhat excessive even for the advanced ham station, but a persistent amateur group has been able to establish "touch-and-go" communication via over-dense meteor trails. Usually the received signals are of very short duration, with only an occasional letter or two of Morse code getting through. Every now and then a large meteor will form a trail of ionization lasting 10 to 20 seconds, or even longer, permitting a burst of intelligence to be transmitted via meteor reflection. During a meteor shower the information rate will increase to the point where a fairly good exchange of intelligence is possible.

Since meteoric ionization takes place in the E-layer, optimum skip distance is from 600 to about 1200 miles. This is just about the same skip that occurs for Sporadic-E reflection and ionospheric scatter. By raising the angle of radiation of the transmitting and receiving antennas it is possible to work within the 600 mile distance because the ionized columns produce a true reflection effect, rather than refraction. Figure 9 graphically indicates the optimum radiation angle for meteoric communication at various distances.

The attenuation vs. frequency curve for meteor trail propagation follows approximately the same course as that for ionospheric scatter, and the graph of Figure 8 will apply to a close approximation for under-dense meteors. It appears, though, that reflections from over-dense meteor trails do not drop off quite so quickly with frequency as indicated by the graph curve, and that these reflections will be detectable well into the VHF region.

Moon Reflection

Perhaps the most amazing demonstration of progress in the communications field is that of detecting VHF signals transmitted to the moon and reflected back to earth. Although it is true that earth-bound scatter circuits are overcoming greater path attenuation than that of the EME (Earth-Moon-Earth) circuit, they do not fire the imagination quite as much as the

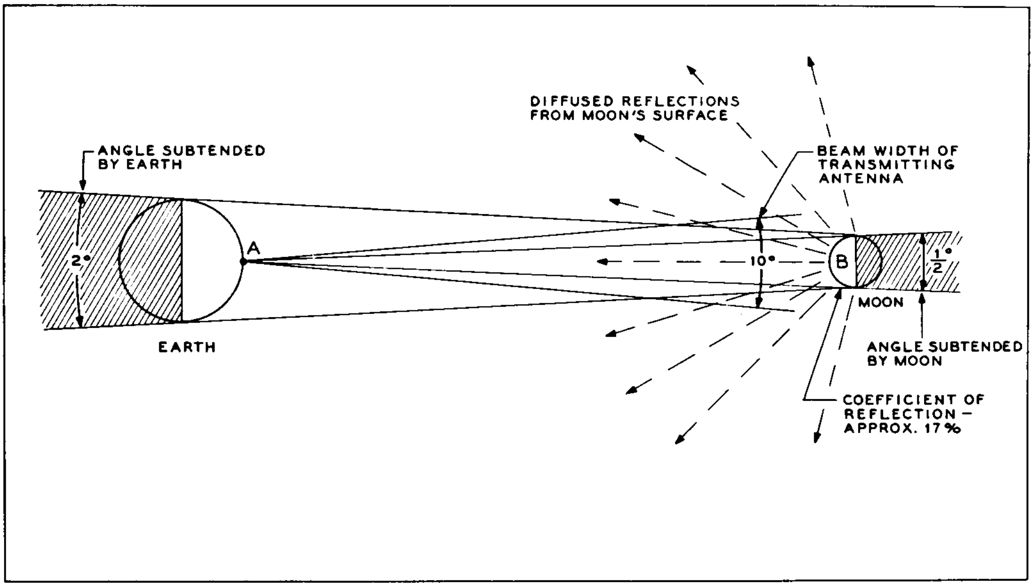


Fig. 18 Earth-Moon-Earth path shows small amount of energy returned to earth.

realization that, in utilizing moon reflection, man is taking his first small step into space.

The basic EME circuit is shown in Figure 18. This drawing helps to show the tremendous obstacles which would at first seem to make detection of moon-reflected signals highly improbable. The earth's satellite has a diameter of approximately 2160 miles and circles the earth at a distance which varies from 221,463 to 252,710 miles. It takes the moon about 28 days to complete one trip around the earth, but we see it move across the sky once each day because of the rotation of the earth.

Although the moon appears very big and bright to the observer, especially when it is full and the sky is clear, it actually subtends an angle of only one-half degree as viewed from the earth. With practical antenna systems having beam widths considerably broader than this figure, most of the radiated energy does not even strike the moon, but passes right by it and into the depths of space. As if this were not a big enough penalty, the part of the signal energy which does strike the moon is largely absorbed. It is estimated that only 17% of the striking energy is reflected, and this amount instead of returning to earth in some sort of beam is actually sprayed all over the heavens. The earth subtends only two degrees when viewed from the moon, so it is quite easy to see that only a minute portion of the energy which left the transmitter is reflected back to the surface of the earth. In addition, the pickup area of the earth facing the moon is 8,000 miles wide, and a large receiving array covers an area of perhaps only 1600 square feet or so. In spite of these staggering penalties to success, the EME circuit can be made to function with equipment well within the capabilities of the radio amateur.

Figure 19 is a plot of EME attenuation in decibels for various frequencies and antenna sizes. Once the antenna size is determined and the frequency of the EME circuit is chosen, this graph will indicate the required power spread between the transmitter and receiver for this particular circuit. The graph of Figure 9 in Chapter 2 will prove helpful in determining bandwidth and transmitter power for the EME circuit.

The EME Circuit

Signal components which are reflected from the center of the moon's disc will return to the earth in a shorter period of time than signals reflected from the edge of the disc. This condition exists in the case of any reflection from a rough, spherical surface, but since the moon has a radius of about 1080 miles the difference in time delay between the two extremes becomes appreciable, and causes a severe reduction in the information rate which an EME circuit can handle. Voice communication cannot be used at all. Slow speed teletype or hand operated keying is about the fastest rate which the narrow bandwidth of a circuit of this type can allow. Exception to this statement may be made if we assume a transmitting antenna with a beam width of ¼-degree, or less, since then we could utilize only the center portion of the face of the moon for reflection. At the present state of the art, such an antenna is impractical.

Ground Reflection Gain

If the transmitting antenna is aimed at the moon when it is near the horizon, ground reflection in front of the antenna will enhance the antenna gain by as much as 6 db. This effect may be compared to that of seeing the image of the sun as it is setting over a body of water. When both the sun and the reflected image are seen, twice as much light reaches the eye as that received from the sun alone. If it is arranged that both the transmitting and receiving antennas of the EME circuit can "shoot" the moon on the horizon the net circuit gain can reach a possible 12 db. The exact amount of surface gain is a function of the ground conductivity in the vicinity of the antennas. When the EME stations are located on a north-south line it is possible for both to utilize the benefits of ground reflection at moonrise and moonset. On the other hand, if the stations have any east-west separation, then only one station at a time can obtain the benefits of ground reflection.

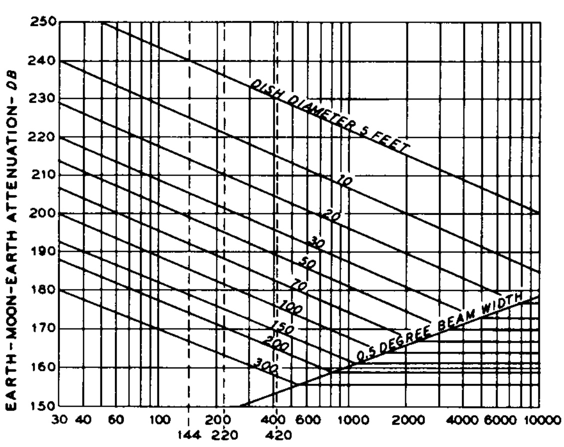


Fig. 19 E-M-E attenuation increases with frequency. At 50 and 144 mc, power required for moon echoes is well within amateur capabilities.

The westerly station will have the benefit during its period of moonrise, and the easterly station will obtain the extra path gain during his moonset. The exception to this occurs when both stations are on exact opposite sides of the earth, in which case they can contact only during the brief time that the moon touches the horizon, setting for one station and rising for the other. Both stations may then utilize ground reflection gain on the EME path.

Most experiments to date have been conducted close to the horizon, taking advantage of ground reflection. However, results have generally been subject to wide fluctuations in signal strength, and it is likely that atmospheric refraction is responsible. Aiming at the moon when it is at some angle above the horizon would tend to reduce atmospheric effects, and would therefore tend to compensate for the loss of ground reflection gain. In other words, if the EME system is designed with sufficient power and antenna gain so that ground reflection is not required, greater signal stability will result.

The Practical EME Circuit

Because the moon may be moving toward or away from the EME stations at speeds up to 980 miles per hour, Doppler shift will change the received frequency according to the formula:

MAXIMUM DOPPLER SHIFT = $2.966 \times f$ (*MC.*)
IN CYCLES (Δf), *AS*
MEASURED AT EQUATOR

Actually, of course, it is the EME stations that are doing most of the moving as the earth rotates. When the moon is rising the Doppler effect increases the received frequency; at moonset the frequency is decreased.

Originally it was thought that the moon might prove to be of great value as a VHF reflector for TV relays, etc. But tests and analysis indicate that its usefulness is quite limited. Not only are bandwidth restrictions quite

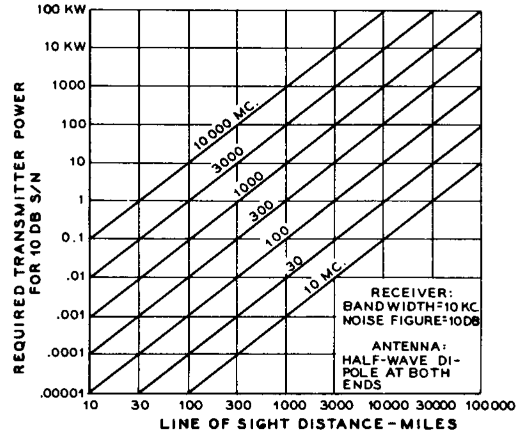


Fig. 20 Attenuation of free space under given conditions increases with frequency as the wavelength factor is important parameter. This paramount fact is discussed in Antenna Chapter of this book.

severe, but the moon is in the sky for only half the time. Therefore, investigation so far has proven of great academic value, and certainly future communication in space will have benefit from the work, but at the present time no commercial application of the EME circuit is in sight.

However, once again the radio amateur may step into the problem and examine it. Moon reflection is entirely within practical limits for the advanced experimenter. The propagation summary at the end of this chapter lists approximate circuit requirements for the EME circuit on various VHF bands.

In order to accurately determine the moon's position as it sweeps across the sky, it is necessary to either delve rather deeply into the study of astronomy, or to engage the services of an astronomer. It will be found that the time of rising and setting changes considerably from day to day, and also that the azimuth position on the horizon during rising and setting varies greatly from true east and west. However, it is not essential that exact data be available. If the sky is clear the antenna system may be "aimed" optically. A copy of the "Old Farmer's Almanac", which may be purchased at many news stands will furnish information on the time of moonrise and moonset, as well as the constellation in which it appears. A "Starfinder" computer (C. S. Hammond & Co., Inc., New York, N. Y.) will also assist in determining the moon's position in the sky during different times of the year. These inexpensive computers are available in book or map stores, and along with the Almanac, will be useful when the sky is overcast, or when the moon is in the sky during the day and is too close to the sun's position to be visible.

Free Space

Quite often the term *free space* is encountered in VHF literature. This is an abstract thought today, but someday we might suddenly find ourselves

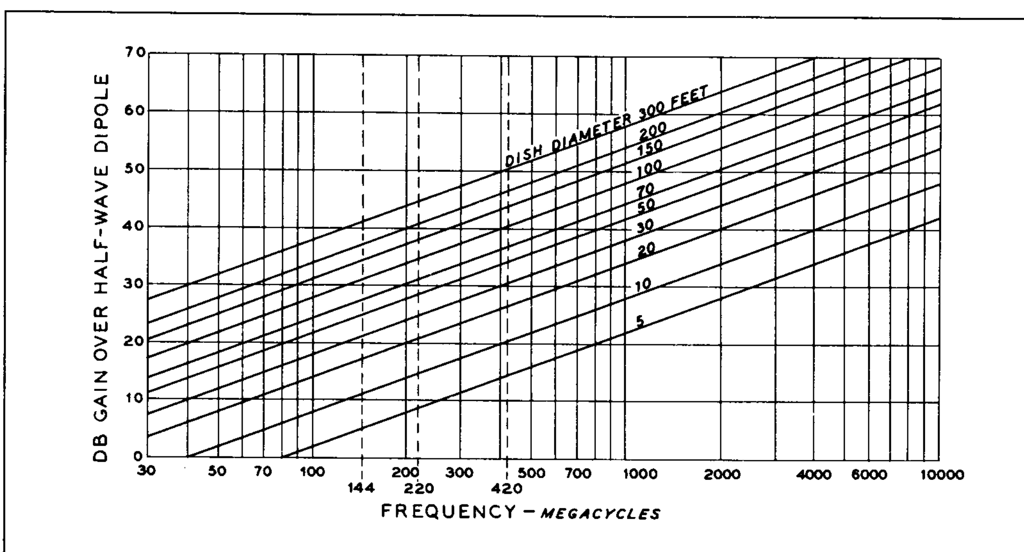


Fig. 21 Parabolic antenna can provide many decibels of gain on long path.

faced with the need for free space communication circuits. The graph of Figure 20 is included from an academic viewpoint—but who can say how many years it will be before actual space ship-to-earth communication is required? The graph illustrates the distances that may be covered at various power levels, assuming fixed receiver bandwidth, noise figure, and antenna gains. Changing any of these factors will modify power requirements accordingly. As can be seen, communication to distances of many thousands of miles with presently available equipment is entirely practical. Indeed, the QRM level on the moon from 2 meter and 6 meter earth stations must be at a high level today! The chaos in the VHF television channels to a casual listener on the moon must give such a legendary figure pause for concern at just what is happening on the planet earth!

The use of parabolic antennas on a free space communication circuit can afford a circuit improvement of many decibels, as shown in Figure 21. High gain antennas and the use of the maximum power limit in the VHF region assure that the radio ham is ready for interplanetary communication as soon as someone conquers the problem of transporting the necessary radio equipment through space!

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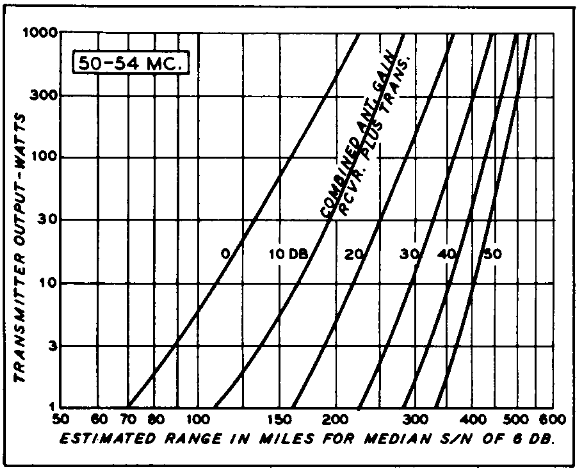
50-54 MC.
SUMMARY OF TRANSMISSION PATH CHARACTERISTICS

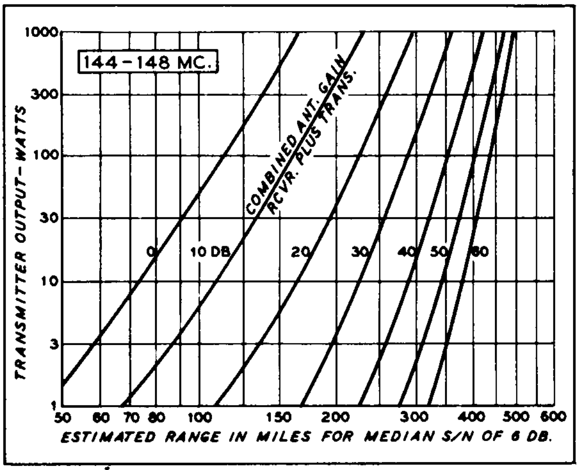


GROUND WAVE AND TROPOSPHERIC SCATTER RANGE

ILLUSTRATING TRANSMITTER POWER IN WATTS VS. RANGE IN MILES. THE GRAPH ASSUMES A RECEIVER NOISE FIGURE OF 2 DB, AND BANDWIDTH OF 3 KC. ESTIMATED APERTURE COUPLING LOSS IS INCLUDED FOR LARGE ANTENNAS, ASSUMING THAT COMBINED GAIN IS ACHIEVED WITH SIMILAR ANTENNAS AT EACH END. THE GRAPH IS BASED ON A SMOOTH TERRAIN. HILLS WILL INTRODUCE SOME DIFFERENCES.

MODE	RANGE (*APPROX. MILES*)	POWER OUTPUT *REQUIREMENTS*	COMMENTS	
SPORADIC-E (*PHONE*)	SINGLE-HOP 600-1400 MILES. DOUBLE-HOP UP TO APPROX. 2500 MILES,	LOW- 5 WATTS OFTEN ADEQUATE.	PEAKS WITH SUNSPOT CYCLE, IN SPRING AND EARLY SUMMER. MAY OCCUR DAY OR NIGHT, BUT TENDS TO PEAK 4 HOURS AFTER SUNRISE AND JUST AFTER SUNSET. POSSIBLE CORRELATION WITH AURORA.	
F_2 SKIP (*PHONE*)	SINGLE-HOP 2500 MILES. MULTI-HOP 12 000 MILES.	LOW-TO MEDIUM 50 WATTS OR MORE.	PEAKS WITH SUNSPOT CYCLE. PEAKS IN WINTER, BUT SPRING AND FALL BEST FOR CROSSING EQUATOR. PEAKS DURING DAYLIGHT HOURS.	
AURORA (*PHONE BADLY GARBLED*)	UP TO 1000 MILES EAST-WEST	MEDIUM 100 WATTS OR MORE. 6 DB ANTENNA.	PEAKS WITH SOLAR STORMS. OFTEN FOLLOWS AN *SID* BY A DAY. PEAKS AT SUNDOWN AND 2 TO 3 A.M. USUALLY LIMITED TO NORTHERN STATES. CHECK *WWV* FOR STORM WARNINGS. POINT ANTENNA NORTH.	
IONOSPHERIC SCATTER (*CW*)	600-1400 MILES	600 W. OR MORE. AT LEAST 12 DB ANTENNA.	PEAKS DURING SUMMER MONTHS. FADES RAPIDLY OVER 20 DB RANGE. NORTH-SOUTH PATHS STRONGER THAN EAST-WEST. PEAKS FROM 6 A.M. TO NOON.	
METEOR REFLECTION (*CW*)	APPROXIMATELY SAME AS IONOSPHERIC SCATTER EXCEPT FOR SPORADIC PEAKS FROM LARGE METEORS, INCREASING DURING SHOWERS.			
MOON REFLECTION (*CW*)	EARTH-MOON-EARTH. APPROXIMATELY 500,000 MILES	600 WATTS INTO ANTENNA	RECEIVER BANDWIDTH (*CYCLES*)	*ANT. GAIN NECESSARY FOR *S/N* OF 4 DB.
			100	17 DB
			400	20 DB
			1000	22 DB
			4000	25 DB
			**RECEIVER NF OF 2 DB. GROUND REFLECTION NOT INCLUDED. ANTENNA GAIN SAME AT BOTH ENDS OF CIRCUIT.*	

144–148 MC.
SUMMARY OF TRANSMISSION PATH CHARACTERISTICS



GROUND WAVE AND TROPOSPHERIC SCATTER RANGE

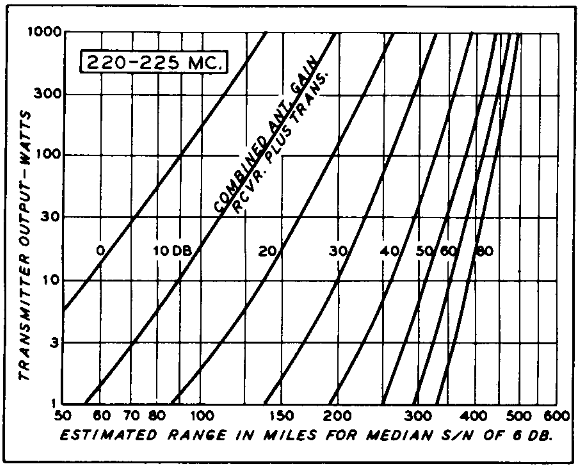
ILLUSTRATING TRANSMITTER POWER IN WATTS VS. RANGE IN MILES. THE GRAPH ASSUMES A RECEIVER NOISE FIGURE OF 3 DB AND BANDWIDTH OF 3 KC. ESTIMATED APERTURE COUPLING LOSS IS INCLUDED FOR LARGE ANTENNAS, ASSUMING THAT COMBINED GAIN IS ACHIEVED WITH SIMILAR ANTENNAS AT EACH END. THE GRAPH IS BASED ON SMOOTH TERRAIN.

MODE	RANGE (*APPROX. MILES*)	POWER OUTPUT *REQUIREMENTS*	COMMENTS
SPORADIC-E (*PHONE*)	SINGLE-HOP 600–1400 MILES DOUBLE-HOP UP TO APPROX. 2500 MILES.	LOW 5 WATTS OFTEN ADEQUATE.	VERY SELDOM WILL SPORADIC-E REFLECT 144 MC. WATCH 50 MC. FOR STRONG OPENINGS. TRANSMIT SIMULTANEOUSLY ON BOTH BANDS IF POSSIBLE.
F_2 SKIP (*PHONE*)	SINGLE-HOP 2500 MILES. MULTI-HOP 12000 MILES	LOW TO MEDIUM 50 WATTS OR MORE.	*MUF* MAY NEVER REACH 144 MC. WATCH 50 MC. FOR STRONG OPENINGS. TRANSMIT SIMULTANEOUSLY ON BOTH BANDS IF POSSIBLE.
AURORA (*PHONE USUALLY UNREADABLE*)	UP TO 1000 MILES EAST-WEST	MEDIUM 100 WATTS OR MORE 10 DB ANTENNA	PEAKS WITH SOLAR STORMS. OFTEN FOLLOWS AN *SID* BY A DAY. PEAKS AT SUNDOWN AND 2 TO 3 A.M. USUALLY LIMITED TO NORTHERN STATES. CHECK *WWV* FOR STORM WARNINGS. POINT ANTENNA NORTH.
IONOSPHERIC SCATTER (*CW*)	600–1400 MILES	IMPRACTICAL WITH AMATEUR POWER LIMITS.	
METEOR REFLECTION (*CW*)	600–1400 MILES	600 WATTS OR MORE. AT LEAST 16 DB ANTENNA.	REFLECTIONS FROM LARGER "OVERDENSE" METEORS ONLY. PEAKS DURING SUMMER MONTHS AND DURING METEOR SHOWERS. MAXIMUM ACTIVITY ABOUT 6 A.M. BURSTS OF SHORT DURATION. HIGH SPEED CW BEST METHOD.
MOON REFLECTION (*CW*)	EARTH-MOON-EARTH. APPROXIMATELY 500,000 MILES	600 WATTS INTO ANTENNA.	see below

RECEIVER BANDWIDTH (*CYCLES*)	*ANT. GAIN NECESSARY FOR *S/N* OF 4 DB.
100	22 DB
400	25 DB
1000	27 DB
4000	30 DB

***** *RECEIVER N/F OF 3 DB. GROUND REFLECTION NOT INCLUDED. ANTENNA GAIN SAME AT BOTH ENDS OF CIRCUIT.*

220–225 MC.
SUMMARY OF TRANSMISSION PATH CHARACTERISTICS

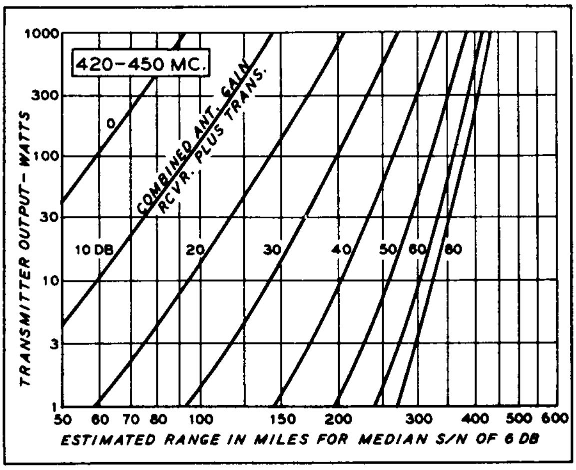


GROUND WAVE AND TROPOSPHERIC SCATTER RANGE

ILLUSTRATING TRANSMITTER POWER IN WATTS VS. RANGE IN MILES. THE GRAPH ASSUMES A RECEIVER NOISE FIGURE OF 4 DB AND BANDWIDTH OF 3 KC. ESTIMATED APERTURE COUPLING LOSS IS INCLUDED FOR LARGE ANTENNAS, ASSUMING THAT COMBINED GAIN IS ACHIEVED WITH SIMILAR ANTENNAS AT EACH END. THE GRAPH IS BASED ON SMOOTH TERRAIN.

<table>
<tr><th>MODE</th><th>RANGE
(APPROX. MILES)</th><th>POWER OUTPUT
REQUIREMENTS</th><th colspan="2">COMMENTS</th></tr>
<tr><td>SPORADIC-E</td><td>NIL</td><td>—</td><td colspan="2">OCCURENCE NEVER REPORTED.</td></tr>
<tr><td>F2 SKIP</td><td>NIL</td><td>—</td><td colspan="2">OCCURENCE NEVER REPORTED.</td></tr>
<tr><td>AURORA
(PHONE UNINTELLIGIBLE)</td><td colspan="2">APPROX. SAME AS 144 MC.</td><td colspan="2">SOMEWHAT GREATER ATTENUATION THAN 144 MC.</td></tr>
<tr><td>IONOSPHERIC SCATTER
(CW)</td><td>NIL</td><td>—</td><td colspan="2">EXCESSIVE ATTENUATION.</td></tr>
<tr><td>METEOR REFLECTION
(CW)</td><td colspan="2">APPROX. SAME AS 144 MC.</td><td colspan="2">—</td></tr>
<tr><td rowspan="6">MOON REFLECTION
(CW)</td><td rowspan="6">EARTH–MOON–EARTH.
APPROXIMATELY 500 000 MILES</td><td rowspan="6">600 WATTS INTO ANTENNA</td><td>RECEIVER BANDWIDTH CYCLES</td><td>* ANT. GAIN NECESSARY FOR S/N OF 4 DB.</td></tr>
<tr><td>100</td><td>24 DB</td></tr>
<tr><td>400</td><td>27 DB</td></tr>
<tr><td>1000</td><td>29 DB</td></tr>
<tr><td>4000</td><td>32 DB</td></tr>
<tr><td colspan="2">* RECEIVER N/F OF 4 DB. GROUND REFLECTION NOT INCLUDED. ANTENNA GAIN SAME AT BOTH ENDS OF CIRCUIT.</td></tr>
</table>

420–450 MC.
SUMMARY OF TRANSMISSION PATH CHARACTERISTICS



GROUND WAVE AND TROPOSPHERIC SCATTER RANGE

ILLUSTRATING TRANSMITTER POWER IN WATTS VS. RANGE IN MILES. THE GRAPH ASSUMES A RECEIVER NOISE FIGURE OF 7 DB AND BANDWIDTH OF 3 KC. ESTIMATED APERTURE COUPLING LOSS IS INCLUDED FOR LARGE ANTENNAS, ASSUMING THAT COMBINED GAIN IS ACHIEVED WITH SIMILAR ANTENNAS AT EACH END. THE GRAPH IS BASED ON A SMOOTH TERRAIN.

MODE	RANGE (*APPROX. MILES*)	POWER OUTPUT *REQUIREMENTS*	COMMENTS
SPORADIC-E	NIL	—	OCCURENCE NEVER REPORTED.
F_2 SKIP	NIL		OCCURENCE NEVER REPORTED.
AURORA	QUESTIONABLE	—	ATTENUATION HIGHER THAN 220 MC.
IONOSPHERIC SCATTER	NIL	—	EXCESSIVE ATTENUATION
METEOR REFLECTION	NOT PRACTICAL WITH 50 WATT POWER LIMIT.		
MOON REFLECTION	NOT PRACTICAL WITH 50 WATT POWER LIMIT.		

CHAPTER IV

VHF Component Considerations

The successful operation of VHF equipment is a function of the degree of integration between the mechanical and electrical designs of the circuit. Great emphasis must be placed upon the correct use of components, their physical size, and their placement in relation to other parts of the equipment. The interplay between electrical and mechanical considerations has as much effect upon the performance of the VHF unit as does the electrical value of the most critical components.

The Wavelength Factor

It must be realized that as one advances into the VHF region the physical dimension of an electrical half wavelength of radio energy begins to assume the proportions of some of the circuit components that make up the VHF equipment. A half wavelength in free space at 50 mc is about 118 inches. At 144 mc a half wave is only 41 inches long, shrinking to 27½ inches at 220 mc. In the amateur 420 mc band, a free space half wavelength measures only 14 inches. At the latter frequency, common interconnecting leads may become resonant antenna systems, and simple parts become huge in terms of the comparatively small wave. A one watt resistor is about .04 wavelength long, and a 6L6 tube is .15 wavelength tall! Odd effects take place when the radio wave approaches in size the components that are supposed to generate, amplify, and receive it, and the small size of the VHF waves has a profound effect upon the electronic hardware to be used at these frequencies. Amateurs used to working and thinking in terms of low frequency equipment are familiar with resistors, capacitors, inductors, and conductors that are small in size compared to the physical dimension of the wavelength handled by the equipment. At frequencies below about 30 mc, small lengths of wire may be employed as interconnections between parts of the r-f circuit and normally cause no trouble, since these bits of copper are short when compared to the wavelength at which the equipment

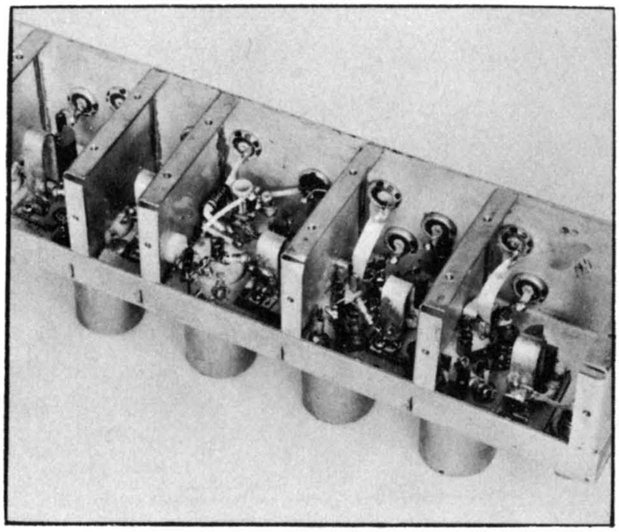


Fig. 1 Specialized design and miniaturized components are required for successful VHF equipment. Shown at right are r-f stages for 420 mc superhet receiver. Note that low inductance leads are made of wide copper strap.

is used. At VHF however, normal physical dimensions of radio components, coils, capacitors, tubes, leads, chassis—all the parts that go to make up the VHF units—begin to approach in size a large fraction of the wavelength. Even the electron becomes relatively cumbersome, as the minute time that it takes to pass from cathode to plate in the vacuum tube becomes an appreciable part of the radio cycle. A metal chassis can no longer be thought of as a fixed point of ground reference, since points of voltage and current maxima may be found along the surface. Metal enclosures may exhibit peculiar characteristics as they cease to be shields and become resonant cavities. Short wires, normally docile at the lower frequencies can turn into radiating antennas .that couple VHF energy into adjacent circuitry. Coils may act as capacitors; capacitors as coils, or perhaps cease to function at all. As the operating frequency is raised, common vacuum tubes lose their efficiency bit by bit until at some critical frequency they cease to function, paralyzed by the minute time lag of electron passage.

VHF Techniques

To the uninitiated, the VHF region is a topsy-turvy world wherein all the normal laws and truths of electronics seem to be continually violated. A study of the details of circuitry and component design in this part of the radio spectrum shows, however, that the natural laws that function so clearly at the lower frequencies still hold true in the world of VHF. The bewildering phenomena noticed at the higher frequencies are observable in smaller magnitude at the lower frequencies where they pose only marginal problems that may often be ignored. Only when the size of the wave begins to approach the size of the circuit components are these natural qualities of the components such a major factor in equipment design.

Rarely do VHF circuit diagrams spell out these problems, leaving the path of the VHF enthusiast strewn with pitfalls and obstacles. Parasitic loops, interlocking ground currents, transit time, distributed capacity, lead inductance, and energy radiation all exist at the lower frequencies, but only at the higher portion of the spectrum do their effects assume such robust

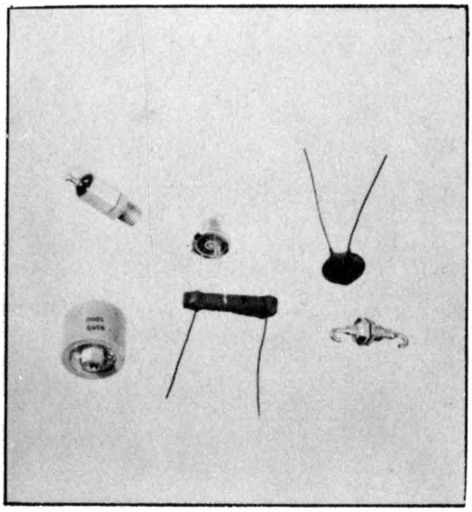


Fig. 2 Specially designed VHF units such as these capacitors exhibit low self-inductance. Button-mica unit at center will perform well into the UHF spectrum. Disc capacitor at left is satisfactory for VHF use.

proportions, and only at these frequencies does their cure and elimination come into such forceful prominence. As a result, the necessity for close correlation between the electrical design and the mechanical properties of the VHF circuit cannot be overemphasized.

The solution to this particular problem is to employ components of the smallest possible size, having a very minimum of the unwanted properties. A new field of design has evolved to overcome the problem of producing components that perform efficiently at very high frequencies. Minute condensers having low inductance leads have been developed, as well as multilead vacuum tubes incorporating close element spacing. An example of application engineering for very high frequencies is shown in Figure 1, and some special VHF components are shown in Figure 2. As can be seen, a process of intelligent miniaturization is the correct approach to the enigma of equipment operation in the VHF region. As the physical size of the radio wave decreases the size of the components handling this wave must also decrease, so that a rough state of equilibrium is maintained between the two. A twilight region is finally reached, bordering upon the UHF portion of the spectrum wherein modified low frequency techniques must be rejected entirely in favor of new concepts in wave motion and equipment design.

Vacuum Tube Limitations

A combination of crippling factors is at work in the common vacuum tube which tend to diminish the efficiency of the tube as the operating frequency is raised. In old style tubes a noticeable decline in operation may be observed as low as 20 or 30 mc. Some common tubes of improved design function well at 150 mc or so, and others operate at 500 mc and higher. Tubes operating in the UHF region bear little or no resemblance to the low frequency counterparts.

At very high frequencies the common triode tube (Figure 3) must be viewed as an a-c circuit element having capacities $C_{g\text{-}p}$, $C_{g\text{-}k}$, and $C_{p\text{-}k}$, and inductances L_p, L_g, and L_k which are inherent in the structure of the tube. These values exist because the tube must have finite size, and must

have sufficient length of leads to permit connection to the circuit components. All of these internal parameters have an important effect upon the operation of the tube.

At low frequencies the external circuit capacitances and inductances are large in value compared to those of the tube, and the effect of the internal parameters is negligible. As the operating frequency of the tube is raised the interelectrode capacitances become a proportionately greater fraction of the total circuit capacitance. At the same time the inductance of the tube leads becomes increasingly important. A point is finally reached where the total resonant circuit may be expressed in terms of the residual parameters of the tube. This frequency is termed the *self-resonant frequency* of the tube.

The inductance of a short tube lead may be surprisingly high in the VHF region. For example, a lead four inches long and .04 inches diameter has an inductance of about 0.1 microhenry. At 1000 kilocycles the impedance of this lead is only 0.63 ohm. At a frequency of 100 mc, the impedance of this short bit of wire is 63 ohms, since the impedance is directly proportional to frequency. An impedance of 0.63 ohms may have no effect upon circuit operation at 1000 kilocycles; an impedance of 63 ohms at 100 mc may very well cause a VHF circuit to cease functioning. To reduce the impedance to the value obtained at 1000 kilocycles, the wire must be reduced in length to less than one-half inch.

Lead Inductance

Inductance of the grid lead of a vacuum tube acts so as to reduce the signal voltage reaching the grid of the tube, as compared to the voltage impressed at the grid terminal. This reduction becomes more prominent as the frequency of operation of the tube is raised. Cathode lead inductance furnishes a common impedance between the grid and plate circuits, introducing undesired coupling, as shown in Figure 4. Plate lead inductance acts so as to reduce the output signal voltage of the tube.

The upper limit of operation of the common vacuum tube is somewhat lower than the self-resonant frequency of the elements, since this limit is also a function of the input loading, the transit time, and the resonant frequency of the L-C circuit.

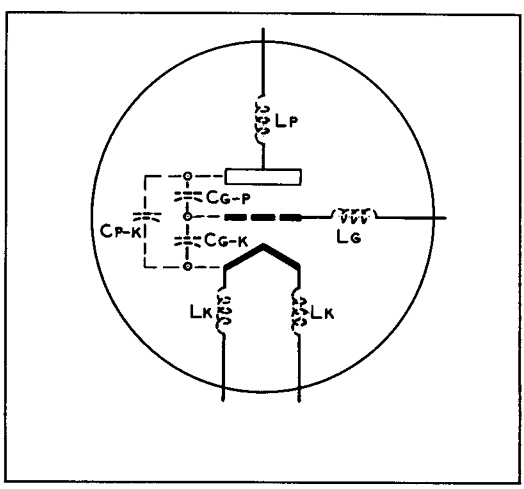


Fig. 3 Simple vacuum tube becomes a complex circuit at VHF. Inductance of leads and internal capacitances combine to increase the internal losses of the tube, as shown.

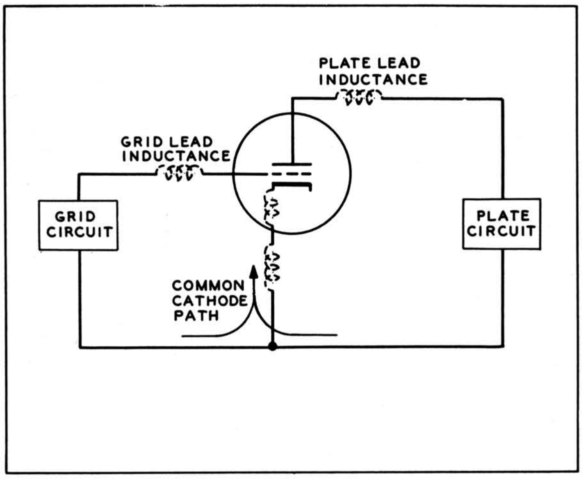


Fig. 4 Cathode lead inductance forms common impedance between input and output circuits. In certain VHF tubes this problem is alleviated by the use of multiple cathode pins.

Lead inductance may be reduced by the use of short, heavy tube leads. In some VHF tubes, multiple leads are employed to further drop lead inductance. The tube elements are mounted close to the glass seal of the tube and in some cases the tube pins are gold plated to reduce the effects of atmospheric corrosion. A cut-away view of a typical VHF tube is shown in Figure 5.

Interelectrode Capacitance

Low interelectrode capacitance may be attained either by reducing the electrode size, or by spreading the electrodes farther apart. Unless abnormally high electrode voltages can be used, separation of the electrodes will have the undesired effect of increasing the electron transit time of the tube. If all the linear dimensions of a vacuum tube are reduced by a factor N, the amplifying characteristics of the tube remain unchanged, but the interelectrode capacitance, lead inductance, and transit time are reduced by the factor N. The upper frequency of operation will be raised accordingly.

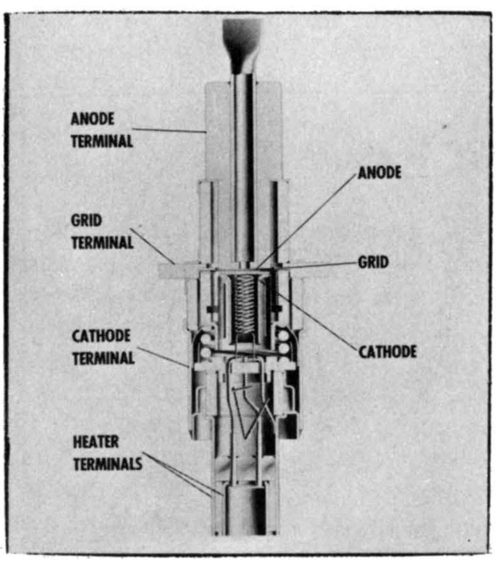


Fig. 5 Radical new design is employed in new "lighthouse" triode of General Electric Co. (GL-6422). Exceptionally strong grid flange, coplanar design and the use of ceramic has resulted in VHF tube that withstands shocks to 400 G's.

However, reduction of the physical size of the tube reduces its power handling ability, since only small areas are present for the dissipation of heat.

Special element configurations have been developed which allow the vacuum tube to become an integral part of a VHF circuit. This is done by virtually eliminating wire leads, replacing them with larger rods or discs which connect the tube element directly to the circuit. The familiar "light house" tubes were forerunners of this technique. Today there are many types of VHF tubes for receiving and transmitting applications based upon this design technique. Improved heat conduction in these tubes allows greater element dissipation and high operating temperatures. Forced air or water cooling is easily applied to transmitting types. One of the more successful very high frequency planar transmitting tubes (GL-6283) is shown in Figure 6.

Electron Transit Time

Under conditions of low frequency operation of a vacuum tube it is usually assumed that electrons leaving the cathode reach the plate of the tube instantaneously. Although nothing in nature happens instantaneously, no harm is done by this assumption as long as the actual time of passage between cathode and plate is negligible compared to the duration of one radio cycle. For example, transit time of 1/1000 microsecond (10^{-9} second) is only 1/1000 of a cycle at a frequency of 1 mc. The same transit time becomes 1/10 of a radio cycle if the operating frequency is 100 mc. It has been found by experiment that a total transit time of 1/10 cycle or less between cathode and plate will permit satisfactory operation of a vacuum tube. At transit times longer than this the tube efficiency drops considerably. When the transit time approaches a quarter cycle the tube will not oscillate. This is caused in part by the phase shift between the plate current and the grid voltage, and in part by the decrease of the effective resistance between grid and cathode resulting from the finite time of electron passage.

For highest oscillator efficiency the electron current should be in phase with the grid voltage and 180 degrees out of phase with the plate voltage. When the electron transit time becomes an appreciable fraction of a radio

Fig. 6 The General Electric GL-6283 is a 250-watt forced-air cooled power tetrode capable of delivering full output up to 900 mc. A coaxial structure is employed, the tube plugging into the end of a coaxial tank circuit. Height of tube is 4¼".

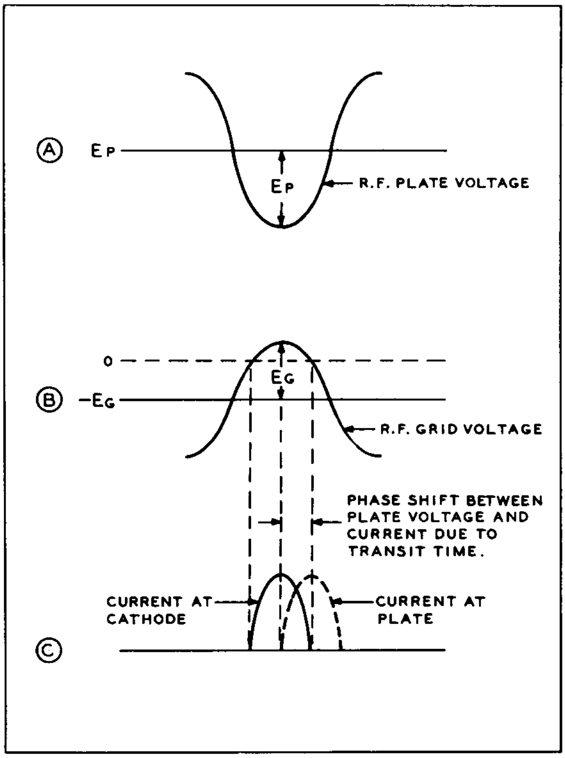


Fig. 7 Electron transit time causes phase shift between plate and cathode current, reducing operating efficiency. Effects of transit time may be decreased by reducing spacing between the electrodes, or by increasing the electrode voltage.

cycle this relation holds only at the instant that electrons are emitted from the cathode. Thus the solid curve of Figure 7C shows the space current that leaves the cathode. This current is in phase with the grid voltage (Figure 7B) and 180 degrees out of phase with the plate voltage (Figure 7A). However, since it takes a fraction of a radio cycle for the electrons to travel from cathode to plate, the current arriving at the plate at any instant must be different from the current leaving the cathode at that same instant. If a maximum number of electrons start toward the plate at the same time they will not arrive at the plate until a short, finite time later. The plate current (shown by the dashed curve of Figure 7C) lags the grid voltage by some small angle and the phase difference between the plate current and the plate voltage is greater than 180 degrees. As a result of this shift, the power output of the tube decreases and the plate dissipation increases. Phase shift within the tube increases in proportion to the square of the operating frequency.

The effects of transit time may be decreased by reducing the spacing between electrodes, or by increasing the electrode voltages. Since insulation considerations prevent raising voltages in many applications, tubes of special design are usually employed where transit time effects may become serious. In many radar applications it is feasible to employ plate voltages two or three times greater than the manufacturer's maximum rating because the unusual way in which the tubes are operated prevents the average element dissipation from exceeding a safe value. The tubes are biased to cut off and plate current is only allowed to flow in short pulses, usually a few microseconds in length. Averaged over a period of time, the dissipation of the elements of the tube is held within normal limits, while exceedingly high power levels may be obtained during the short pulse periods.

Input Loading

When a negatively charged electron approaches an electrode it induces a positive charge on the electrode. As the electron converges upon the electrode the positive charge flows to the electrode. As the electron recedes, the positive charge flows away from the electrode. In this fashion the electrons that form the plate current of a vacuum tube cause electrostatically induced currents in the grid structure as they move past it. In a low frequency oscillator circuit where the grid is negative and the transit time negligible, the number of electrons approaching the grid is always equal to the number of electrons going away from the grid. The current induced on one side of the grid by the approaching electrons is equal to that induced on the opposite side by the receding electrons. Since these induced currents are in opposition the net effect is zero.

If the transit time of the oscillator tube is an appreciable part of a radio cycle the number of electrons approaching the grid is not equal at all times to the number going away. As a result, the electrostatically induced currents do not cancel. Thus, grid current can flow in a VHF oscillator even when the grid is at a negative potential relative to the cathode. This current consists of a movement of positive charges back and forth in the grid structure and the effect is to produce a loss which is equal to that taking place in an imaginary input resistor connected between grid and cathode of the tube. The value of this resistor decreases rapidly as the operating frequency of the tube rises. In the very high frequency region this resistance may become so low that the grid is practically short circuited to the cathode, preventing proper excitation of the tube. The effect of this loss is to raise the operating temperature of the grid structure, imposing another limiting factor upon the high frequency operation of the tube.

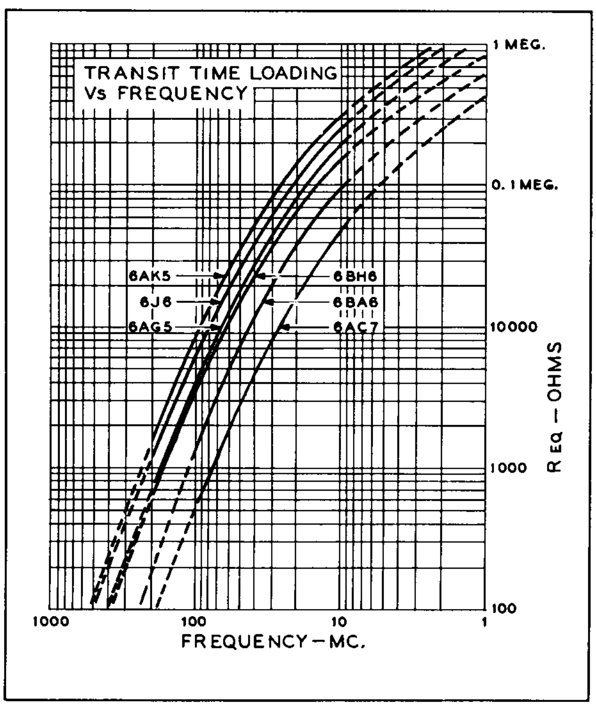


Fig. 8 Transit time loading increases rapidly with frequency. Input resistance of most common tubes drops to unuseable value above 100 mc or so, as shown in chart.

It should also be noted that the input capacitance of a tube will vary as the induced grid charge of the tube changes. The input capacitance is also a function of the cathode lead inductance and the transconductance of the tube.

Tube losses caused by the combination of transit time, induced grid charge, and input loading effects may be expressed in terms of the *hot input conductance* of the tube. The input impedance of the tube when no cathode current is flowing is referred to as the *cold input conductance,* whose components are the reactance of the input capacitance and a residual resistance caused by dielectric loss in the insulating material within the tube. The total input circuit loading of a vacuum tube is a function of both the hot and the cold conductance of the tube.

By these subtile variations the parameters of a vacuum tube change as the frequency of operation is increased and, as may be expected, the changes in no way enhance the operating efficiency of the tube. In spite of these handicaps, modern advances in the art of designing and building tubes, spurred by the expanding use of the VHF region for television and point to point communication have supplied the amateur with a wealth of tubes capable of first grade performance at these frequencies.

Circuit Noise

Absolute receiver sensitivity in the very high frequency region is not determined by atmospheric noise as it is at the lower frequencies. It is rather a function of the magnitude of circuit noise of the receiver which is composed of antenna noise, input circuit noise and tube noise.

Circuit noise is masked at the lower frequencies by the high external noise picked up by the antenna. This external noise is composed of atmospheric noise generated by static electricity, cosmic noise received from outer space, and various kinds and intensities of man-made noise. The level of external noise seems to drop off gradually at higher frequencies until in the 100 mc region it is quite low. Circuit noise, on the other hand, is of minor importance until a frequency of 50 or 70 mc is reached. At this point the external noise level becomes quite low, and the circuit noise of the receiver starts to form the "floor," or minimum level beneath which it is difficult to receive signals. By proper design, the circuit noise can be reduced to a great degree, permitting reception of extremely weak VHF signals. To achieve these results, it is necessary to understand the circuit problems associated with various types of receiver noise.

Thermal Agitation

In any substance the molecules are always in a state of random motion. Temperature, in fact, is a measure of this random motion, and it is only at absolute zero (-273 degrees Centigrade) that such motion ceases. As the temperature of the substance rises above absolute zero the random motion of the molecules increases. The motion of the molecules in a conductor varies the instantaneous position of the free electrons of that conductor, and this agitation of the electrons corresponds to a minute electrical current flowing in the conductor and is known as *thermal agitation*

noise, or *Johnson Noise.* This noise is a random fluctuation and is generated over a wide band of frequencies. (An analogy may be taken from the study of white light which is composed of all the colors of the spectrum. Random noise is often termed *white noise.*)

When this noise is observed in any receiving system the noise occupied parts of the spectrum falling outside of the passband of the receiver will not contribute to the noise output of the receiver. Only the noise received within the receiver passband is of consequence. Thus the first step to obtain a minimum amount of thermal noise is to limit the bandwidth of the receiving system to only that amount required to accept the intelligence passed by the communication circuit, and no more. The magnitude of random noise in any given circuit is expressed as a current whose value is:

$$i = \sqrt{\frac{4KT\Delta f}{R}} \qquad (1)$$

K = BOLTZMAN CONSTANT (1.38×10^{-23})
T = TEMPERATURE (°K)
Δf = BANDWIDTH (CPS)
R = RESISTANCE OF CIRCUIT (*OHMS*)

The bandwidth for general purposes is taken at the half power points, or 0.7 voltage points of the selectivity curve of the receiver.

Antenna Noise

At the terminals of any antenna appears a random voltage equivalent to the thermal noise generated in a resistance equal to the radiation resistance of the antenna. This noise is caused by electron agitation in the elements of the antenna, and is a function of temperature. A theoretically perfect "noiseless" receiver connected to an antenna would therefore have a noise output with which the weakest received signal would have to compete. In passing, it is interesting to note that added to this antenna noise may be a varying degree of galactic noise received from outer space. This latter noise will vary in intensity depending upon the heading of the antenna, the time of day, and various subtle changes in the ionosphere.

Receiver Noise

Agitation noise exists in all parts of the receiver, competing with the signal applied to the input circuit. The greatest portion of random noise in the receiver is generated in the tubes, particularly those in the input stages of the receiver. Because of the many factors that determine the excellence of a VHF receiver, it has been found necessary to establish a yardstick, or *figure of merit* by which receiver operation in this part of the radio spectrum may be judged. The term *noise figure* is used to express the ratio of the noise power of any receiver compared to that of a perfect receiver. The noise figure (N/F) is actually a measure of how nearly a receiver approaches the theoretical minimum of residual noise, and is expressed in decibels. A "noiseless" receiver would have a N/F of zero db. Noise figures of 2 to 3 db have been obtained in the region of 150 mc, rising to perhaps 6 db at 400 mc, and 10 to 12 db in the microwave region. Noise figure is a summation of many sources of noise, all referred to the grid of the input stage

of the receiver. Some of these sources are: 1—Thermal noise of the receiver input circuit, 2—Random noise of the first r-f tube, and 3—Noise contributed by the second r-f tube. The total noise contributed by all these sources is the square root of the sum of the squares of each individual noise source.

Tube Noise

The tube selected for the input stage of a VHF receiver will to a large extent determine the overall performance of the receiver. The most important factor to be judged in the choice of this tube is the amount of random noise it produces. A measure of this noise may be expressed as the *equivalent noise resistance* (R_{eq}) of the tube. This term indicates the value of an imaginary resistance whose thermal noise is equal to the tube noise. The value of R_{eq} is expressed in ohms, and this term may be used to calculate the generated noise voltage for any given bandwidth of the amplifier stage by using this formula:

$$E = \sqrt{1.6 \times 10^{-20} \, \Delta f \times R_{EQ}} \qquad (2)$$

E MEASURED IN MICROVOLTS AT ROOM TEMPERATURE (20° C)

When the R_{eq} of two different tubes is known their relative noise figure of merit is known, regardless of the operational bandwidth of the stage, whereas the operating bandwidth enters into calculation of the noise voltage of the stage. R_{eq} is a function of the noise generated within the tube

TUBE TYPE	CIRCUIT	GM	REQ	TUBE TYPE	CIRCUIT	GM	REQ
6J6	T	5300	470	6CB6	P	6200	1440
6J4	T	11000	230	6AC7	P	9000	720
12AT7	T	6600	380	6AN4	TM	2500	1600
6BK7	T	6100	410	6J4	TM	2750	1450
6BQ7	T	6400	390	6J6	TM	1575	2540
6BZ7	T	6800	370	2C51	TM	1375	2900
6AN4	T	10000	250	12AT7	TM	1650	2430
417-A	T	24000	105	5687	TM	2500	1600
5687	T	10000	250	6AG5	PM	1250	6600
2C51	T	5500	455	6AK5	PM	1280	7520
6AC7	T	11000	220	6BZ6	PM	1525	5840
6AK5	P	5100	1880	6U8	PM	1300	9120
6AU6	P	5200	2660	6X8	PM	2100	7780
6BA6	P	4400	3520	6BA7	PENTAGRID MIXER	950	61700
6BZ6	P	6100	1460	6SA7	PENTAGRID MIXER	450	240000
6AH6	P	9000	720	6SB7Y	PENTAGRID MIXER	950	61700

T = TRIODE AMPLIFIER　　TM = TRIODE MIXER

P = PENTODE AMPLIFIER　　PM = PENTODE MIXER

Fig. 9 Equivalent Noise Resistance of common tubes is shown in above cnart. High transconductance and low interelectrode capacitance make good VHF tube.

itself, which is composed mainly of *shot noise* and (in the case of multigrid tubes) *partition noise.* Shot noise is caused by random variations in the plate current of the tube. When amplified, this noise sounds like a shower of shot falling upon a metal plate, hence the name. Partition noise is caused by a random division of the cathode current between the various elements of a multielement tube. A third type of noise is *induced grid noise* caused by fluctuations in the cathode current passing the grid. The effect of this latter noise may be modified by the design of the input circuit of the tube.

The magnitude of R_{eq} (expressed in ohms) for various tubes is:

(3) Triode Amplifiers: $$R_{EQ} = \frac{2.5}{G_M}$$

(4) Triode Mixers: $$R_{EQ} = \frac{16}{G_M}$$

WHERE:
I_P = PLATE CURRENT
I_{SC} = SCREEN CURRENT
G_M = TRANSCONDUCTANCE
I_K = CATHODE CURRENT

(5) Pentode Amplifiers: $$R_{EQ} = \frac{I_P}{I_P + I_{SC}}\left(\frac{2.5}{G_M} + \frac{20 I_{SC}}{G_M{}^2}\right)$$

(6) Pentode Mixers: $$R_{EQ} = \frac{I_P}{I_P + I_{SC}}\left(\frac{16}{G_M} + \frac{20 I_{SC}}{\left(\frac{G_M}{4}\right)^2}\right)$$

(7) Multigrid Mixers: $$R_{EQ} = 20\,\frac{I_P (I_K - I_{SC})}{I_K \left(\frac{G_M}{4}\right)^2}$$

A low value of R_{eq} is desired for tubes operating in the VHF region. It can be seen from these formulas that, in general, triode tubes should outperform pentode tubes, and that any type of mixer tube has a relatively poor equivalent noise resistance. A table of R_{eq} for some of the more common tubes is given in Figure 9. It can be observed that partition noise is a relatively large factor in determining R_{eq}. Mixer tubes drawing large values of screen current have a very high partition noise, and a correspondingly high value of R_{eq}.

There seem to be wide variations in the equivalent noise resistance of individual tubes of the same type. In general, under constant circuit parameters those tubes drawing the largest value of plate current tend to be less noisy. Minute variations in the manufacturing process and small traces of gas within the tube are thought to account for this variation.

A second factor entering into the choice of the input tube is the *transconductance.* This is a measure of the amplification factor and the plate resistance of the tube. To obtain the highest possible power gain in the input stage, the transconductance figure of the tube used in this position must be high. As shown in Figure 9 the higher transconductance tubes tend to have the lowest value of R_{eq}.

A third consideration is that the input and output capacitances of the tube must be low. Measured noise figures for tubes having low interelectrode capacitances tend to be somewhat lower than calculated values. In particular, the 6AK5 has a lower noise figure than the 6AC7 at 50 mc, whereas the opposite conclusion would be reached from inspection of the data of Figure 9.

As the frequency of operation of any given tube is raised the noise figure tends to degenerate. It is interesting to note that the noise figure of pentode amplifier stages deteriorates at a faster rate than does the noise figure of an equivalent triode amplifier. This effect is undoubtedly due to a rapid increase in partition noise in the multielement tubes.

The second stage of the receiver may also contribute in part to the overall noise figure, its contribution to circuit noise being a function of the gain of the first stage, and the R_{eq} of the second stage. In cases where the noise figure and gain of the first r-f stage are marginal the noise figure of the receiver may be dependent upon the noise generated in the tube following the input stage. The use of a high gain, low noise tube in the first stage will reduce the possibility of second stage noise determining the noise figure of the receiver.

Determination of Receiver Noise Figure

The noise figure for any receiver may be computed if certain factors are known. These are: 1—Antenna Radiation Resistance, 2—Antenna transformer tuned impedance, 3—Transit time loading, 4—Equivalent noise resistance of the first r-f tube, and 3—Noise contributed by the second r-f stage. Of these parameters, the transit time loading is almost impossible to calculate. It may be measured, and values of loading for certain tubes are presented in Figure 8.

Noise figures are usually obtained by direct measurements made upon the receiver. The receiver input is terminated with a resistor of low reactance and proper resistance. A random noise, usually generated by thermal agitation in a *diode noise generator,* is injected into the input circuit of the receiver. First, the power output of the receiver is measured with the noise generator output at zero level. Then, the generator output is raised to a level which produces a 3 db increase in receiver noise output. The noise figure of the receiver is a function of the ratio of these two levels, and may be computed from these measurements.

In many cases, determination of the exact noise figure is not necessary since it is only necessary to make comparative checks as the receiver is adjusted for the optimum noise figure. Discussion of noise figure measurements and noise generators for running these tests will be discussed at length in Chapter 12 of this Handbook.

Radiation Loss

Certain effects that may pass unnoticed in the high frequency region become of great importance at very high frequencies since the wavelength is small in comparison to circuit components. One of the first phenomena to be observed is that *energy radiation loss* becomes quite high in this portion of the spectrum. Some amateurs have been mystified at the loss of circuit efficiency in some VHF equipment when shielding was removed from the tuned circuits, little realizing that the r-f field of the components was being coupled to free space and that the drop in efficiency was caused by radiation of energy through the opening in the shield.

The common two wire transmission line used successfully at low fre-

quencies becomes progressively more inefficient as the frequency of operation is raised. Radiation and eddy current losses occur in greater magnitude, as a result of the incomplete cancellation of the electromagnetic field surrounding the wires. If the two parallel conductors of the line are spaced closely, on the order of 1/100 wavelength or less, the field around one conductor almost completely neutralizes the field around the other. In the VHF region the normal line spacing is large compared to the wavelength and an efficiency drop is caused by radiation of energy from the line. A decrease in conductor spacing below a certain minimum value imposes a serious limitation on the maximum voltage that can exist between the conductors without insulation breakdown. The use of coaxial line instead of open wire line almost completely reduces radiation loss because the outer conductor of the line acts as a shield, preventing the electromagnetic field of the inner conductor from escaping into space.

Energy radiation loss acts upon a circuit much in the same manner as additional loading, at the expense of the permissible useful loading. The end result is poor circuit efficiency, lower circuit Q and less useful output. A primary requisite, therefore, of VHF component choice is to make sure that the components are arranged and shielded in such a way so as to reduce radiation loss to an absolute minimum.

Once again the same perplexing enigma created by the physically small size of the VHF wave compared to the operating circuit raises design problems that are peculiar to this portion of the radio spectrum.

Skin Effect

An important factor operating to the detriment of the VHF circuit is *skin effect*. This is the tendency for high frequency currents to flow in a thin layer on the surface of the conductor. The depth of penetration of the current in copper is:

$$d = \frac{2.63 \times 10^{-3}}{\sqrt{f}} \qquad (8)$$

where D is the depth of penetration in inches, and f is the frequency in megacycles. Thus, the higher the frequency, the thinner will be the layer of the conductor in which the current flows. The I^2R losses that take place in a given conductor must also increase as the frequency increases, since the cross section area of current flow is less. Skin effect is reduced by using conductors of larger cross section so that the current can flow through a reasonably large area even though the depth of penetration is small. The use of large leads for VHF circuitry reduces not only the lead inductance but also lead resistance caused by skin effect. A further reduction can be brought about by plating the conductor with a low-resistivity metal, such as silver. In cases where corrosion could convert the surface of the conductor to a highly resistive oxide the conductor may be plated with gold which does not corrode and which has good conductivity.

TABLE OF DISSIPATION FACTORS		
MATERIAL	SOFTENING POINT (Oc)	D/F AT 100 MC.
BAKELITE	50-135	.005-.08
FORMICA	—	.02-.05
KEL-F	—	.003
LUCITE	70	.006
CERAMIC	1400	.001-.003
GLASS	520	.001-.003
TEFLON	66	.0002
POLYETHYLENE	100	.0002
POLYSTYRENE	82	.0001
MICARTA	—	.056
PLEXIGLASS	95	.007
NYLON	65	.02

Fig. 10 Dielectric loss is a measure of energy stored in dielectric of material. Best VHF insulators have low dielectric loss, and also low dissipation factor. Teflon, Polyethylene, and Polystyrene are among the best.

Dielectric Losses

Dielectric Losses in insulating materials are objectionable at VHF because of the accompanying reduction of circuit output and efficiency. The best VHF dielectric is air, and a minimum of solid dielectric insulating material should be used in r-f circuitry. Dielectric loss acts as a *dissipation factor* which is a measure of the r-f energy stored in the dielectric. Dissipation factors of various materials are listed in Figure 10. The best insulators have the smallest dissipation factor, and include certain ceramics, polyethylene, polystyrene and teflon.

Conductors

Soft-drawn, single strand copper wire or strap is, in general, the best material to employ as a conductor in the VHF portion of the spectrum. Stranded wire may be employed where connection to a moving part is required, but the r-f loss in such wire is substantially higher than in an equivalent section of solid wire. Because of skin effect, the high frequency current tends to concentrate near the surface of the wire, particularly at sharp corners. A round conductor, therefore, provides the best balance of current concentration. The self inductance of a single wire at high frequencies is important and this factor must be watched in resonant circuits.

The coaxial transmission line has advantages at these frequencies which make it very practical for amateur use. Standardized coaxial lines were developed during the last few years that incorporate extremely low r-f loss, combined with high transmission efficiency. Of these lines, type RG-8/U is by far the most popular type. This cable has an inner conductor composed of seven strands of #21 AWG copper wire surrounded by an extruded sheath of polyethylene. This material provides good high frequency insulation from the standpoint of loss and flexibility. Teflon insulation is even better, especially in temperature stability, although it is much more costly. The outer conductor of the cable consists of a woven metal braid of #34 copper wire. A protective outer jacket made of polyvinyl chloride protects the cable against mechanical abuse and moisture. The r-f attenuation of RG-8/U line is about 2.1 db per hundred feet at a frequency of 100 mc, rising to

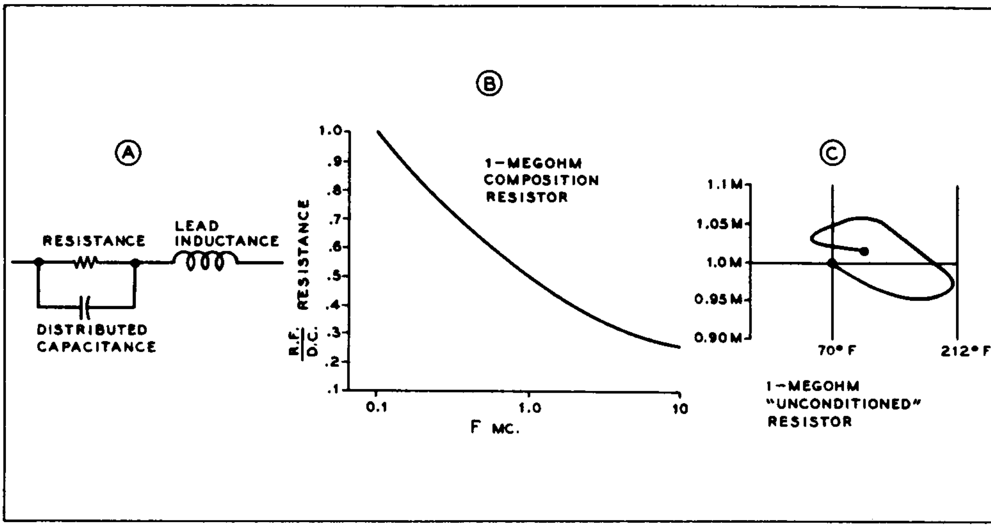


Fig. 11 The simple composition resistor is a complex circuit element at VHF.

about 5 db per hundred feet at 400 mc. If lower transmission line loss than this is required it is necessary to employ more expensive lines having a teflon or air dielectric. In some instances, balanced open wire lines having a conductor spacing of one inch or less have been employed with success at frequencies above 100 mc. With care in construction, the radiation loss of the open wire transmission line may be kept to an unusually low figure at frequencies below 150 mc. Additional information concerning transmission lines is given in Chapter 7 of this Handbook.

Resistors

Composition resistors generate thermal agitation noise, which is a function of the temperature of the resistor, its resistance, and the bandwidth over which the measurement is made. The open circuit noise voltage is given by:

$$E = \sqrt{5.5 \times 10^{-23} T R \Delta f} \qquad (9)$$

WHERE: E MEASURED IN MICROVOLTS
T = TEMPERATURE (°K)
R = RESISTANCE OF CIRCUIT (OHMS)
Δf = BANDWIDTH (CPS)

As an example, a 0.1-megohm resistor at a temperature of 298 degrees Kelvin (25 degrees Centigrade) will generate 2.68 microvolts of noise over a bandwidth of 5 kilocycles.

A second source of noise in composition resistors is *carbon noise* created by random changes in the resistance as current flows through the unit. Johnson noise is relatively constant regardless of frequency, whereas carbon noise tends to decrease with frequency, and also with the IR drop across the resistor. Carbon noise is rarely a cause for concern in the VHF region, as it is masked by the thermal agitation noise of the resistor.

Any resistor of finite size will exhibit residual capacity across its terminals, and capacity to nearby objects. The equivalent circuit of a composition resistor, therefore, must include these values which may be of great impor-

tance in the VHF region. In addition, the resistor has a small value of self inductance, the exact value depending to a degree upon the manufacturing process (Figure 11A). The r-f resistance of a composition resistor tends to drop at the higher frequencies, particularly in the case of high values of resistance. Values of r-f resistance as low as 1/5 the d-c resistance have been measured at frequencies near 50 mc. The drop in r-f resistance is a variable factor, depending upon the size of the unit and the method of manufacture. A typical curve for a ½-watt composition resistor is shown in Figure 11B.

The composition resistor is unusually sensitive to temperature changes (Figure 11C). The heat cycle in most cases is a random one, and the resistor rarely returns to the resistance value at which the cycle started. It is important, therefore, to shield the body of the resistor from excessive heat while it is being soldered into the circuit. The resistor lead to be soldered should be grasped with a long nose pliers before the resistor body, allowing the soldering heat to be absorbed by the pliers without undue heating of the body of the resistor.

The common wire wound resistor may be pictured as a low-Q inductance having distributed capacity as shown in Figure 12A. The resistor forms a self resonant circuit (Figure 12B), the r-f resistance at the resonant frequency approaching several hundred times the value of the d-c resistance. At other frequencies the reactance of the wire wound unit may be either inductive or capacitive. At the parallel resonant frequency the wire wound resistor makes an excellent r-f choke. But, as can be seen from Figure 12B, the unit should be used with caution in r-f circuitry at high frequencies, since its equivalent r-f resistance may bear no relation to the d-c resistance value of the resistor.

Resistors having low values of distributed capacity and reactance have been developed for use in the VHF region. IRC type *F* resistors may be used with good results in the 150 mc region, and the IRC type *MPM* resistors have acceptable characteristics at even higher operating frequencies. Special resistor discs are produced by various manufacturers that may be employed in the microwave region. These discs consist of plastic frames coated with a resistive material. Such resistors are often used to terminate

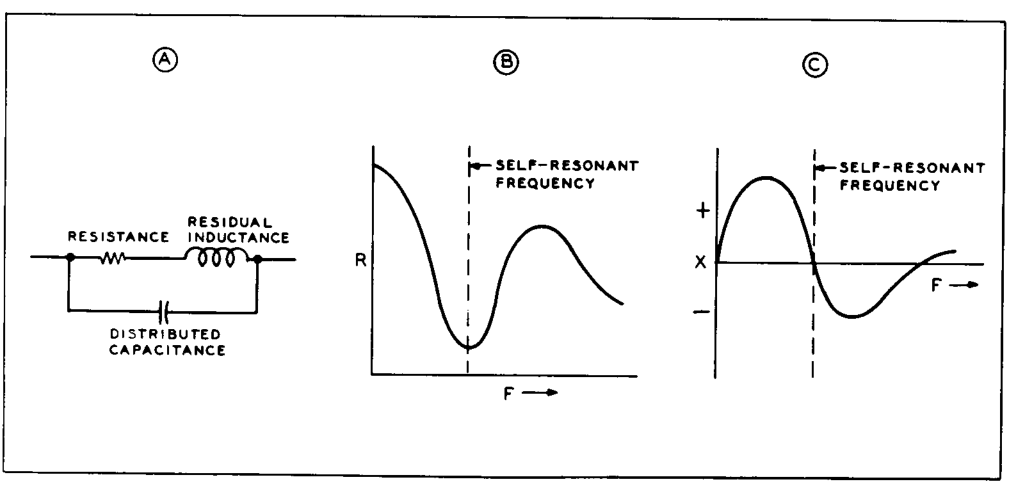


Fig. 12 The wirewound resistor acts as a resonant circuit in the VHF region.

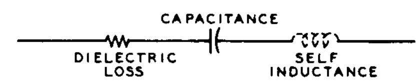


Fig. 13 Capacitor has self-resonant frequency falling in the HF or VHF range. Paper capacitors should not be employed for VHF work. Ceramic, or silver mica units having low internal inductance and short, heavy leads are suitable.

waveguides and UHF transmission lines. These resistors are often built directly into the end of a waveguide, and are water cooled to provide maximum power dissipation.

Capacitors

The equivalent circuit for a capacitor at very high frequencies is given in Figure 13. It is essentially a series resonant circuit, the frequency of resonance being determined by the inherent inductance of the unit. This inductance is a function of the dielectric material of the unit, the size, shape and composition of the plates of the capacitor, the type of leads, and the manufacturing technique. Simple paper capacitors such as are used for bypass purposes in audio equipment and broadcast receivers have a very low series resonance in the range of 3 to 5 mc. Mica dielectric capacitors have less internal inductance than do paper units, having series resonances falling in the vicinity of 12 to 40 mc.

For operation up to and including the 144 mc band the inexpensive disc ceramic capacitor is satisfactory. If the leads of these units are cut as short as possible the .001 ufd units will function satisfactorily up to approximately 170 mc or so. For operation at higher frequencies the "button" mica capacitor is recommended. This unit possesses extremely low lead inductance and may be employed at frequencies well above the 420 mc amateur band. Various other types of capacitors have been designed for VHF use. One of the most popular types is the "Coaxial" unit, shown in Figure 2. The self inductance of such a unit when it is suitably mounted in a grounded plate is extremely low. Standoff capacitors are useful for bypassing socket terminals and tie points, and different types of "feed thru" capacitors may be used to decouple leads passing through shield partitions.

To determine the self resonant frequency of a capacitor the two terminals of the unit should be short circuited by a low inductance strap. A grid-dip oscillator may then be employed to determine the resonant frequency of the closed circuit. Any additional lead length in the circuit will of course lower the natural resonant frequency of the capacitor. A capacitor should not be used at an operating frequency higher than its natural self resonant frequency.

CHAPTER V

VHF Circuitry

The two cornerstones of VHF circuitry design are the use of short, low inductance leads, and physically small components. This is the starting point from which all other design techniques evolve. It is true that a whole new design philosophy is needed in VHF work since not only the components, but the material upon which they are mounted and the space between the components become a vital part of the circuitry. In addition, as has been discussed in the previous chapter all circuit components must be considered as complex items having unwanted but permanent bits of inductance, resistance, and capacity that exert a profound effect upon circuit design and physical layout. If the circuit is designed with these component qualities in mind they may be made to work for, and not against, the overall figure of merit of the VHF equipment.

The Chassis

Either plated steel, aluminum, brass, or copper may be employed as chassis material for VHF equipment. Of the three materials, copper has the lowest r-f resistance per unit area, but is highly susceptible to oxidation and corrosion. In some cases, a copper chassis may be either silver or gold plated to reduce oxidation effects, and then lacquer dipped for maximum serviceability. The high thermal absorption of copper must be taken into account when it is used as chassis material. In some instances, instability in VHF equipment has been traced to poor grounds made to a copper chassis resulting from the inability of the usual soldering iron to heat the chassis sufficiently to prevent the formation of a high resistance joint. It is therefore recommended that thin brass instead of copper be used for small construction work (such as converters, exciters, etc.) when it is desired to solder ground connections directly to the chassis.

For general VHF use the commercial aluminum or plated steel chassis are satisfactory. Care must be taken when ground connections are made to an aluminum chassis, since it is difficult to "fuse" a solder joint through

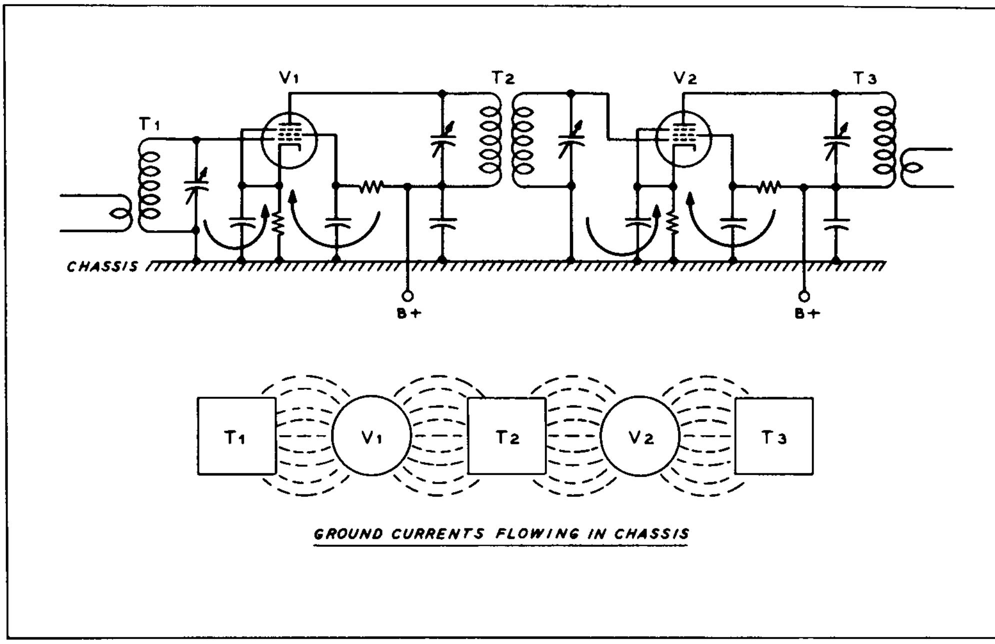


Fig. 1 Grid and plate circuit return currents flow through chassis of stage.

the oxide that coats the material. For best results, all ground connections to aluminum should be of the pressure type. Even so, trouble may eventually develop in such a joint as a result of electrolytic action between the aluminum and a copper wire under damp conditions.

Ground Currents

Each VHF stage that is mounted on a chassis has its own set of r-f ground currents flowing between the input and output circuits. These ground currents are illustrated in Figure 1. To satisfy the condition of highest stability, the ground currents of different stages should not be allowed to couple. Coupling between ground currents of adjacent stages will lead to spurious signal paths, circuit instability and parasitic oscillations.

Ground currents may be restricted to a minimum chassis area by making the physical layout of the stage such that the return leads from the input and output circuits to the cathode of the tube of the stage are as short and direct as possible. Each stage should lie in a straight line, one following the other. If the equipment stages are not in a line, but curve around a corner, as shown in Figure 2, the possibility of coupling between the ground currents of two stages is increased. In this case, additional interstage shielding or a reduction in stage gain is required to maintain the stability of the equipment.

It must be remembered that ground currents flow *on the surface* of the chassis, and not through the metal of the chassis. Trouble will be encountered in some cases when the components of a stage are mounted atop the chassis and the terminals of the tube project beneath the chassis. The ground return path in this case may be some round-about path through a chassis hole near the tube. It may even be the mounting hole of the tube socket. Unwanted coupling between the tube pins and the ground currents flowing

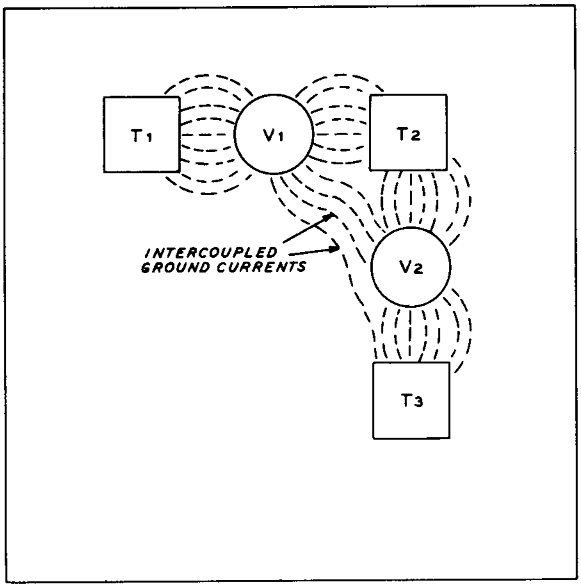


Fig. 2 Intercoupled ground currents result when layout of stages passes around a corner. Instability of amplifier is a result of such inter-coupling.

through the tube socket mounting hole may cause circuit instability. The solution is to mount the components on the same side of the chassis as the tube terminals or to provide separate ground returns *through* the chassis for both the input and output circuits, as illustrated in Figure 3.

R-F and Non-R-F Grounds

It should be noted that there are two different types of grounds in a VHF circuit. *R-F grounds* are those that are concerned with the input and output circuitry of the equipment (Figure 1). These ground circuits carry the r-f currents of the stage and should be as short and direct as possible, having a minimum of intercoupling between them.

Non r-f grounds are illustrated in Figure 4. These grounds are not actually

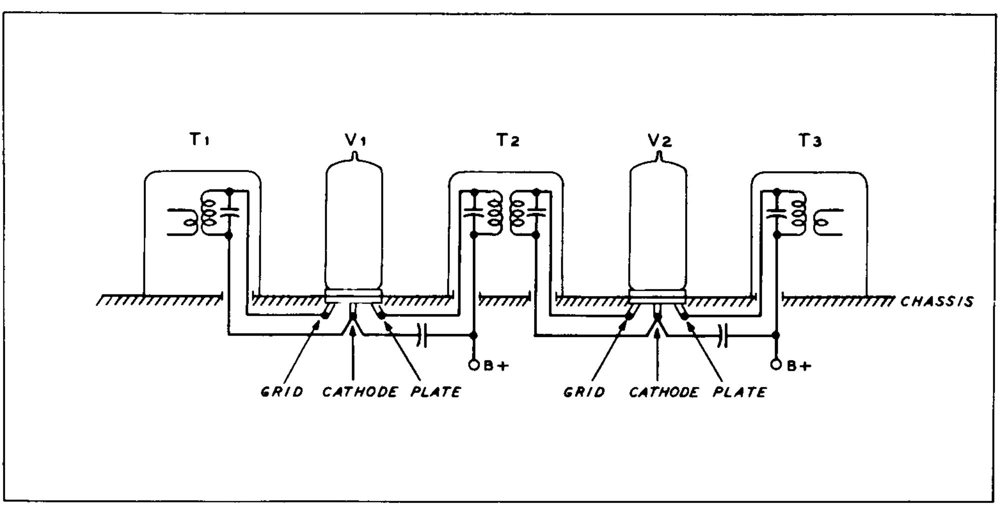


Fig. 3 Separate grid and plate ground returns to cathode reduce possibility of regeneration when tuned circuits (T1, T2, and T3) are above the chassis.

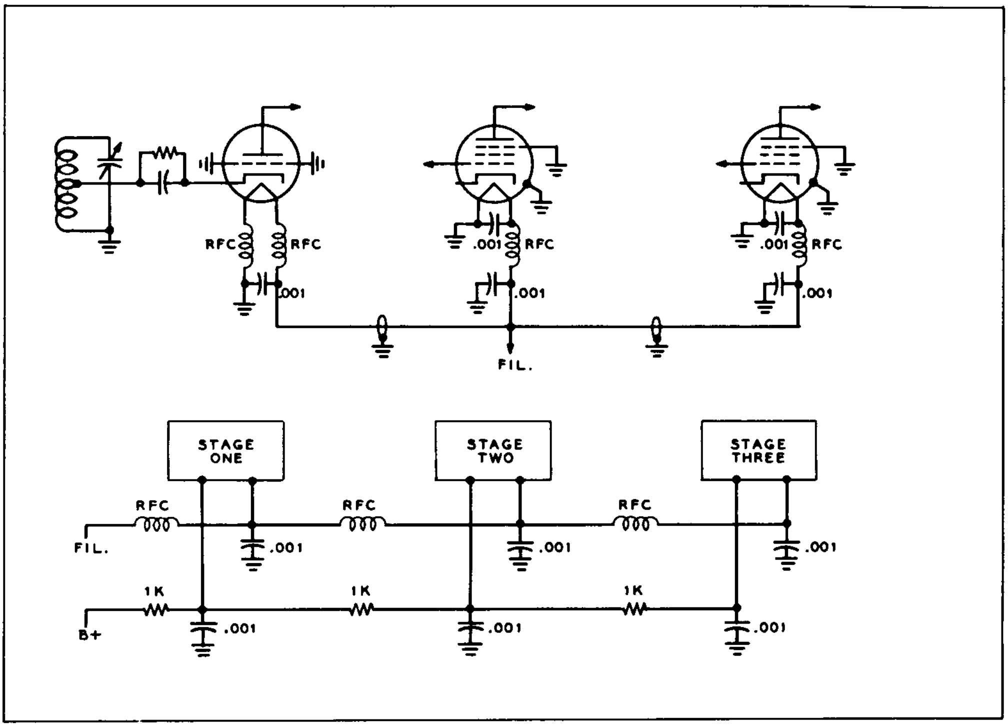


Fig. 4 Decoupling of filament and plate supply leads is important VHF step.

a part of the r-f circuitry of the stage, and may be considered to be d-c circuit grounds. These include one leg of the tube filament, the filament by-pass condenser, the shell of a metal tube, and various d-c return circuits of the stage. Good construction practice calls for these non r-f grounds to be attached to the chassis at a point other than that chosen for the circuit grounds to insure a minimum of coupling between input and output circuits of the stage.

The heater circuit of VHF tubes is especially critical as to grounding techniques. Since the heater of the tube has finite length, and in most instances is capable of being grounded only at one end, the heater within the tube may "rise" above ground potential as a result of the inherent inductance of the element. Minute r-f currents would then flow along the heater, giving rise to unwanted spurious coupling circuits within the stage. Common filament connections will therefore provide coupling between VHF stages unless steps are taken to isolate the interconnecting filament leads. The most common cure is to employ a separate ground for one filament leg, and to bypass the ungrounded filament pin to the same ground point. Series r-f chokes may be placed in the filament leads between tubes, and the filament leads may be shielded, if necessary. In some designs, dual filament chokes are used in each stage, as shown in Figure 4.

Plate and screen voltage supply leads may be decoupled by means of simple R-C or L-C filters, as shown in Figure 4.

Equipment Enclosures

The shield or box in which the VHF equipment is placed may play an

important part in determining the stability of the design. It has been pointed out that extensive shielding of r-f circuitry is necessary in the VHF region to prevent radiation loss and/or intercoupling between adjacent circuits. It must be remembered, however, that any metal enclosure is coupled to the electric and magnetic fields produced inside the shield by the equipment. The enclosure may then act as a waveguide or resonant cavity having attenuation dependent upon the relationship between the size of the enclosure and the operating frequency of the equipment. If the overall gain figure of the equipment placed within an enclosure is greater than the "waveguide" attenuation of the enclosure measured between the input and output circuits of the equipment, circuit instability is sure to be in evidence (Figure 5). Thus, the size and shape of the enclosing shield may play an important role in establishing the circuit stability of the VHF equipment. Regeneration of this type may be spotted easily, as it changes in intensity when one side of the enclosure is removed. The instability can be cured by additional shielding placed around the input stage of the equipment, or by a change in the physical dimensions of the enclosure.

Holes and Leads

It is necessary to provide holes in shield enclosures for the purpose of allowing air circulation for proper ventilation, and also to permit the introduction of power and coupling leads to the apparatus. The use of such openings in "r-f tight" boxes raises the problem of spurious energy coupling through these holes. Electromagnetic energy travels upon the surface of a conductor, and the energy inside an enclosure would have no way of escaping unless there is some sort of a break in the box. In addition, energy existing outside the box can be coupled inside the shielded area through the same openings in the shield.

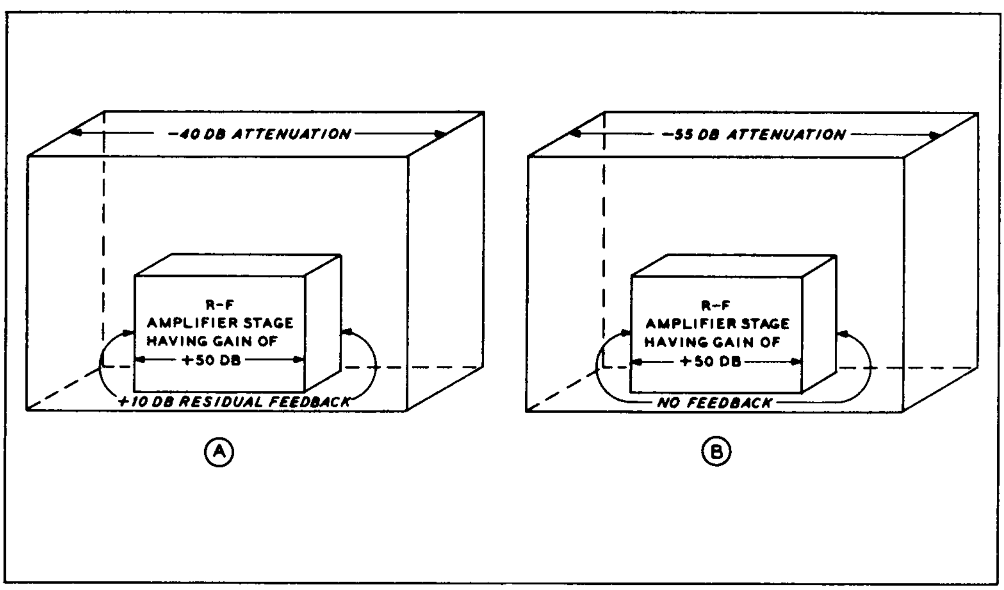


Fig. 5 If the gain figure of equipment within an enclosure is greater than "waveguide" attenuation of enclosure (A), circuit instability may result.

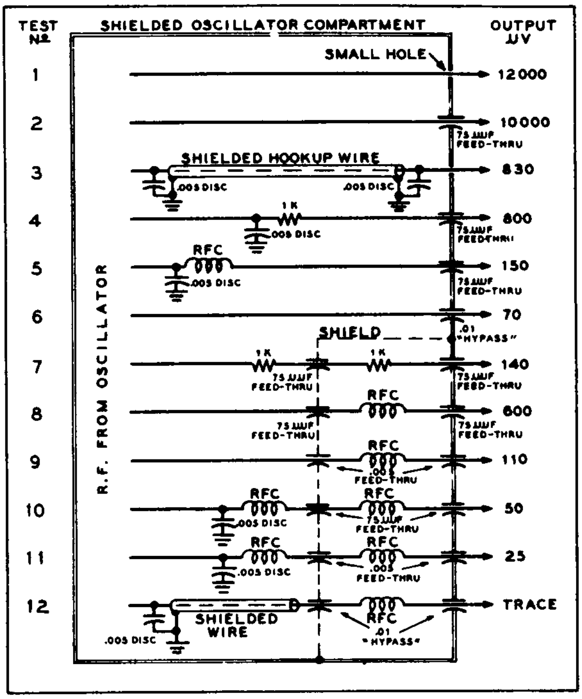


Fig. 6 Careful attention must be paid to lead filtering to achieve highest attenuation. Multiple filtering, as in Test 12 proves most effective.

The simplest opening in an enclosure is a small, round hole. The leakage of energy through such a hole can be measured, and is found to be a function of the radius of the hole, increasing in strength rapidly as the radius of the hole grows larger. Small holes of the order of 1/8-inch diameter or less may be employed in the VHF region in enclosures, allowing equipment ventilation without excessive energy leakage from the enclosure. Mack Seybold, W2RYI reported some extensive shielding experiments in the June, 1949 issue of CQ magazine. He determined that when an oscillator was placed in a solid metal enclosure with no protruding leads or holes, the signal attenuation on the outside of the box was greater than -40 db. A copper screen, dipped in solder to provide a metallic contact at each wire intersection showed the same degree of attenuation in the VHF region when substituted for one side of the box. When the screen was replaced by a perforated plate containing 1/4-inch holes, the attenuation of the box dropped a small, but measureable amount. A hole large enough to mount a meter completely destroyed the effectivity of the shield at 53 mc. A slot in the enclosure caused by imperfect fit of the sides reduced the shielding ability of the box by a large degree.

A slot opening in an "r-f tight" box is equally if not more destructive of the shielding ability than is a hole, since the leakage of the slot is governed in most cases by the longer dimension. The gap acts much in the manner of a *slot antenna*. It is therefore necessary to provide good bonding across all slotted openings to prevent the "antenna effect" from taking place.

When a conductor or lead passes through a hole in an enclosure, the conductor will pick up energy inside the box and radiate it on the outside. It is therefore necessary to establish the conductor at the potential of the shield, both inside and outside the box. A metal tuning shaft, for example, should pass through a bushing which grounds it securely to the wall of the shield. Wire leads should pass through a suitable filter network located inside the enclosure.

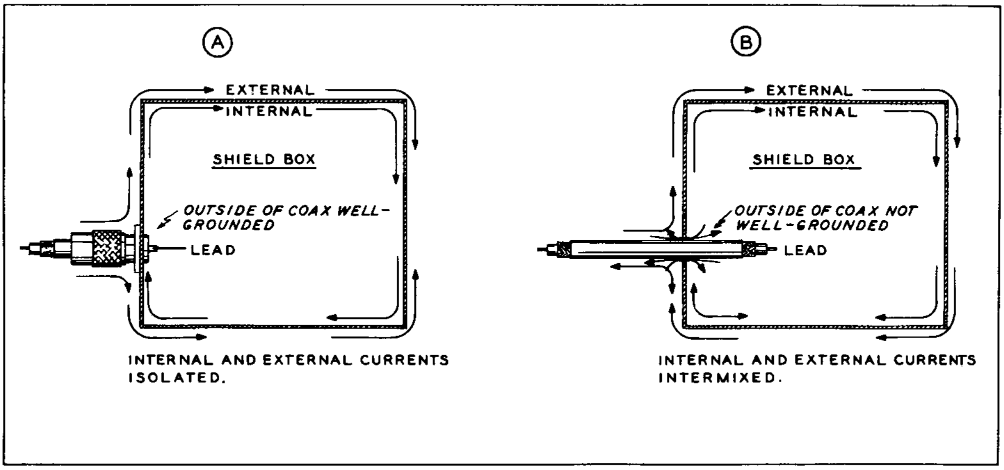


Fig. 7 Coaxial connector must be used to prevent ground current leakage.

Results of experiments with various filter circuits at 50 mc are reported by Phil Rand, W1DBM in his book *Television Interference*, published by the Laboratory of Advanced Research, Sperry-Rand, Inc., South Norwalk, Conn. A summary of these results is tabulated in Figure 6. It can be seen that a combination of an effective filter network plus careful shielding of power leads within the enclosure will reduce energy leakage through the shield to a minute amount.

Attention must also be given to the physical termination of r-f leads that enter the "r-f tight" box. In the VHF range, coaxial line is usually employed for such leads. In order for the coaxial line to function properly, the r-f field must be entirely contained within the outer shield. The outside of the shield must be maintained at ground potential, even though the inside of the shield has r-f current flowing along its length. It is important, therefore, that the shield of the coaxial line make a complete bond to the enclosure at the point of cable entrance, as shown in Figure 7A. If the coaxial line is brought into the box, and the outer shield of the line is grounded at some point *inside* the box the r-f currents normally contained within the enclosure will have an opportunity to escape via the outside of the coaxial shield, as shown in Figure 7B. The effectivity of the enclosure will then be degraded to a considerable degree.

It must be remembered that the amount of shielding and filtering necessary in VHF work depends upon the power gain of the equipment, and the allowable amount of radiation loss that can be tolerated. High gain circuits operating anywhere in the VHF region will require a proportionately greater degree of shielding and filtering between stages than simple, low gain stages. Above fifty mc or so, most r-f equipment requires extensive shielding and filtering, since energy loss from radiation becomes progressively greater as the size of the wavelength approaches the size of the component parts of the unit.

Meter Assemblies

It becomes necessary upon occasion to mount a meter in the shield enclosure of a VHF unit. The two- or three-inch hole required by a meter will cause a discontinuity in the shield and will allow a passage of r-f

energy through the shield. To minimize this leakage, it is necessary to filter and bypass the leads to the meter, and to encase the meter in a metal box that makes electrical contact to the shield entirely around the meter, as shown in Figure 8. A simple meter shield may be cut from the end of a tin can of the correct diameter. The edges of the can are fluted with a tin-snip to make good electrical connection to the enclosure. The meter studs pass through two insulated holes drilled in the "bottom" of the can. A shield of this type may also be used to cover potentiometers or other panel controls that would otherwise permit energy leakage through the shield.

Internal Circuit Leads

Conductors that penetrate internal partitions within an enclosure may contribute leakage through the partition, as shown in Figure 9A. The internal field of one compartment may be picked up by a loop of the wiring and passed through the hole in the shield, to be radiated in the next compartment. The efficiency of such energy transfer is enhanced if one of the transfer loops happens to be self resonant at the operating frequency of the equipment. To eliminate this leakage path it is necessary to by pass the conductor directly at the point it enters the shield hole in such a way as to form a low impedance path to ground for the spurious current flowing in the circuit. Button mica condensers mounted directly in the partition are often employed for this type of supression. In some cases, it is necessary to place shielded r-f chokes on each side of the partition to further reduce signal leakage.

VHF Parasitic Oscillations

Various forms of parasitic oscillations exist in VHF circuits, much as they do at the lower frequencies. These unwanted oscillations are caused by resonant circuits existing in the input and output leads of the amplifier, excited by coupling within the tube. In most cases, the cure for these oscillations is a combination of correct circuit design and the use of attenuation traps at the frequency of oscillation. In order to evaluate the parasitic circuit, it is necessary to examine the neutralizing circuits employed in the VHF region.

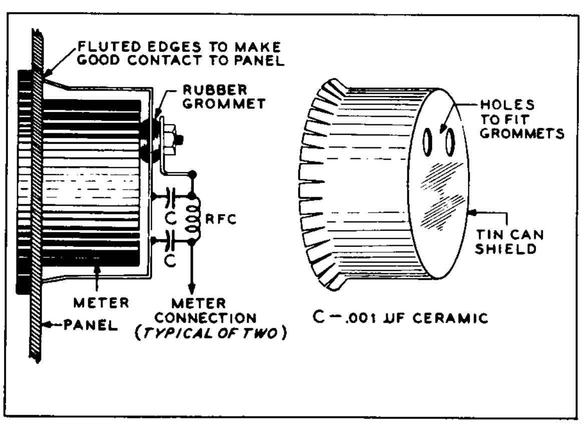


Fig. 8 Simple meter shield may be made from end of tin can. Radiant energy from inside of cabinet will be shielded from meter mounting hole in panel.

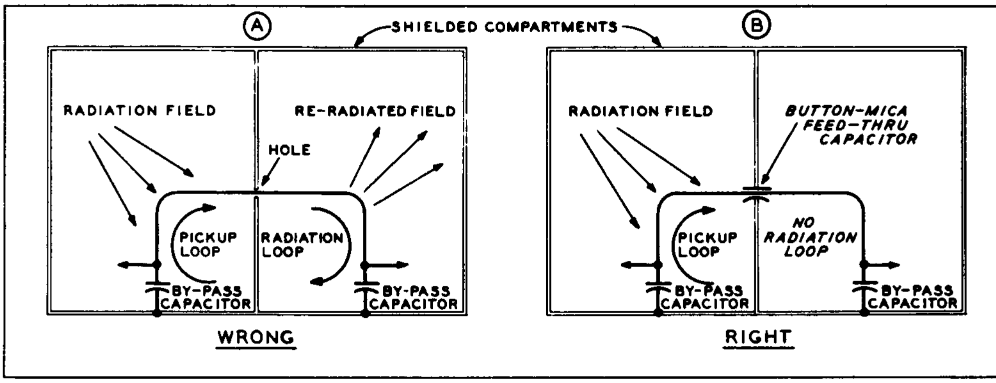


Fig. 9 Energy can be radiated from area by lead through hole in wall of shield. Proper bypassing (B) will cut lead radiation and seal compartment.

Analysis of Neutralizing Circuits

Figure 10 is a drawing of the elements of a tetrode tube operating in the VHF region. The tube elements involved in the feedback circuits are indicated. These circuit elements are inherent and inside the vacuum enclosure of the tube, and involve the residual plate-to-grid capacity, the plate-to-screen capacity, the screen-to-grid capacity, and the inductance of the screen lead of the tube. It will be noted that the r-f voltage developed in the plate circuit (E_P) causes a current I to flow through the plate-screen capacity ($C_{p\text{-}s}$) and the inductance L in the screen lead. The passage of this current through L develops a voltage drop (-E) which has a polarity opposite to that of the plate r-f voltage.

In Figure 11 these circuit elements and voltages have been arranged in graphic form wherein the vertical axis represents the magnitude and polarity of the r-f voltage of that part of the circuit with respect to the cathode r-f potential. Because all of the circuit components involved are pure reactances, the voltages shown are either in-phase or out of phase and so can be represented as positive or negative with respect to each other. The voltages plotted are the components of the r-f output voltage E_P, and no attempt

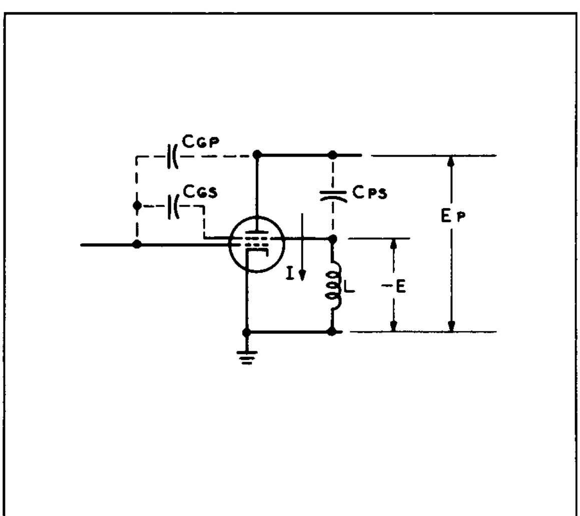


Fig. 10 Residual parameters of tetrode tube in VHF region include grid-plate capacity, grid-screen capacity, plate-screen capacity, and screen lead inductance. Voltage drop across screen inductance is out of phase with plate circuit r-f voltage.

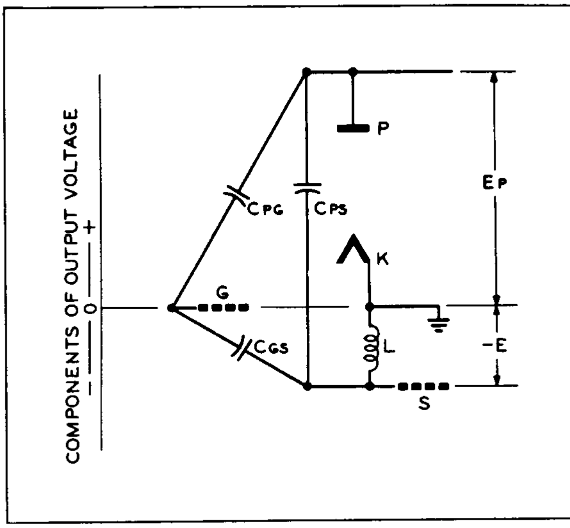


Fig. 11 Circuit elements of tetrode tube are arranged in graphic form, wherein all components are referred to cathode potential. At low frequencies screen inductance is small and screen grid is at cathode potential and normal neutralizing circuits will apply.

is made to show the normal driving voltage on the grid of the tube. The plate P is shown at a high positive potential, represented by the dimension E_P. The voltage developed across the screen lead inductance places the screen at a negative voltage with respect to the r-f plate voltage. The screen of the tube is shown to be below the filament line by the amount -E. If the circuit were perfectly neutralized the control grid would lie on the zero potential axis, since no component of output voltage is then developed between grid and filament.

The total r-f voltage between plate and screen is composed of E_P plus -E. This voltage is applied across a potential divider consisting of the plate-grid capacity ($C_{p\text{-}g}$) in series with the screen-control grid capacity ($C_{g\text{-}s}$). When this capacity divider is suitably matched to the magnitudes of E_P and -E, the control grid will have no voltage difference to cathode resulting from the output circuit potential, E_P. This situation is shown in Figure 11.

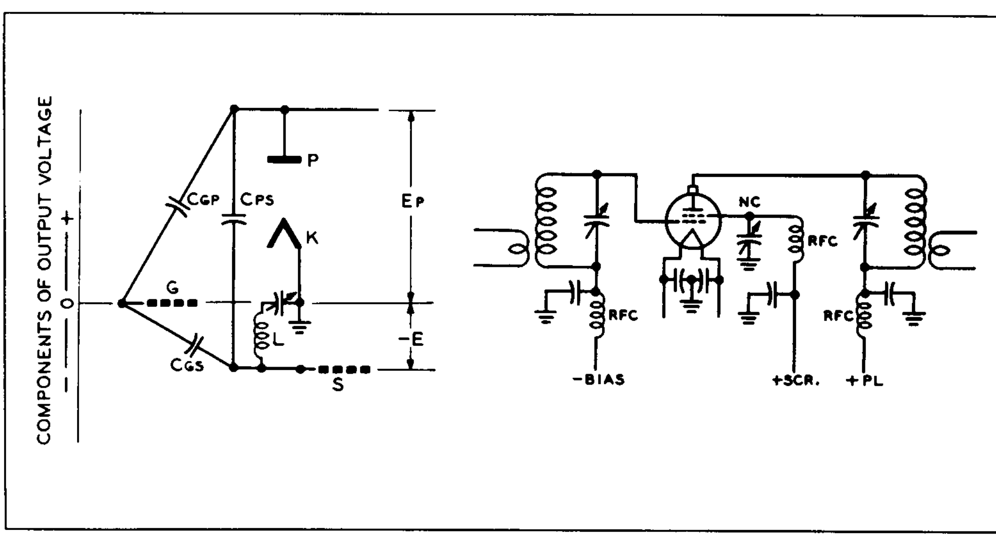


Fig. 12 Series resonant circuit may be used for screen neutralization at VHF. Screen lead inductance is cancelled out by capacitive reactance of NC.

It should be noted that the potential dividing action between $C_{g\text{-}p}$ and $C_{g\text{-}s}$ is not affected by the operating frequency of the tube. On the other hand, the division of voltage between plate and screen, and screen and ground due to the screen charging current I varies greatly with frequency. There will, therefore, be *one* particular frequency at which this capacity dividing network places the grid at the cathode potential as far as plate circuit action is concerned. This frequency is called the *self-neutralizing frequency* of the tetrode. At this one frequency the tube is inherently neutralized due to the circuit elements within the tube structure, and the external inductance of the screen path to ground. The self-neutralizing frequency of older type tetrodes and pentodes may be as low as 20 mc, while tubes of later design have this critical frequency fall in the 50-200 mc region. The self-neutralizing frequency of certain of the new external anode VHF tetrodes falls in the region above 400 mc.

Tetrode Neutralization

When the tetrode is operated below the self-neutralizing frequency the normal neutralizing circuits apply, as shown in Figure 14. If the operating frequency is higher than the self-neutralizing frequency, the voltage -E developed across the screen lead inductance is too large to allow the proper division of voltage between the internal capacities of the tube. One method of reducing the voltage developed across L is to series tune the screen lead to ground, providing a low screen to ground impedance at the operating frequency of the tube, as shown in Figure 12. This form of neutralization is frequency-sensitive and must be readjusted whenever a large change occurs in the operating frequency of the transmitter.

Another method of neutralizing the tetrode in the VHF region is to vary the potential divider network made up of the internal capacities of the tube. This can be done by adding capacity external to the tube between the grid

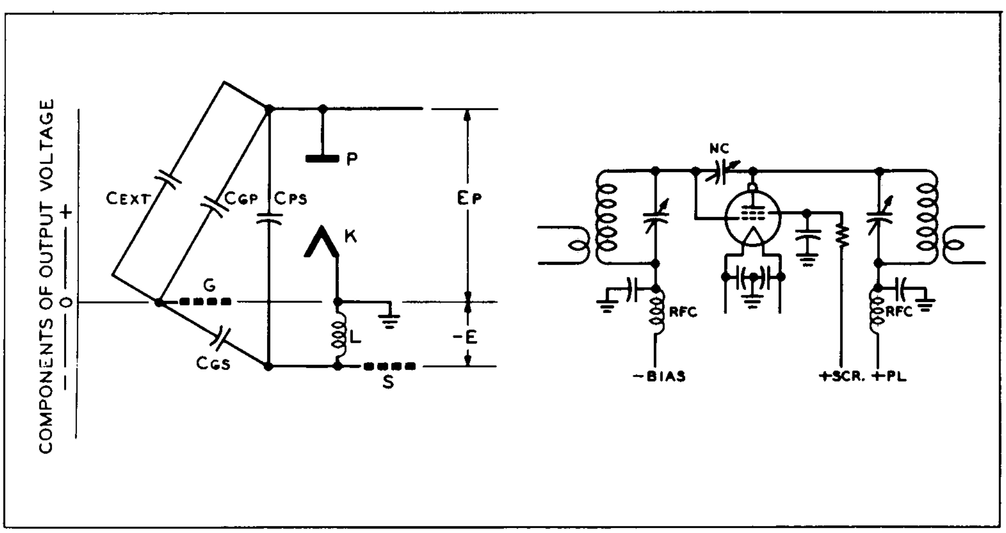


Fig. 13 Capacitor-divider network made of grid-plate capacity of tube plus external capacity, and grid-screen capacity provides neutralization at VHF.

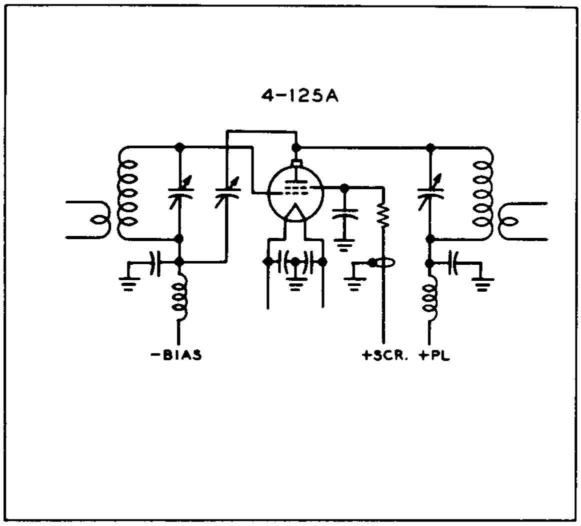


Fig. 14 Bridge neutralizing circuit may be employed at frequencies below the self-neutralizing frequency of tetrode tube. The normal neutralizing procedures apply.

and plate, as shown in Figure 13. The extra capacity is of the same order of size as the residual grid-plate capacity of the tetrode.

If the r-f power amplifier is to operate above the self-neutralizing frequency of the tetrode and is required to tune over a range of frequencies, it is probably easier to employ the screen series tuning neutralizing circuit and to make this control available to the operator. If the operating range includes the self-neutralizing frequency of the tube, this circuit is also desirable because the incidental lead inductance in the neutralizing condenser lowers the apparent self-neutralizing frequency of the circuit so that adjustment of neutralization may be obtained over the desired frequency range. Obviously, if the frequency range is great, switching of neutralizing circuits will be required, especially if the range falls both above and below the self-neutralizing frequency of the tube. In most cases a small 100 mmfd. variable condenser in the screen lead has been found satisfactory for this circuit.

VHF Parasitic Suppression

VHF parasitic oscillations usually take place above the self-neutralizing frequency of the tetrode tube. When the tube is operating in this region, suppression of oscillations may usually be achieved by the correct neutralization of the stage, as explained earlier. However, when the tube operates below the frequency of self-neutralization, it is possible for parasitic oscillations to take place above this critical frequency. The usual neutralizing precautions taken to neutralize the tube at the operating frequency will, in this case, merely increase the degree of r-f feedback at the parasitic frequency. Conversely, parasitic neutralization will tend to increase r-f feedback at the frequency of operation. This occurs since a complete phase shift in the neutralizing circuit occurs as the frequency of self-neutralization of the tube is passed. As an example, let us examine a power amplifier using a 4-125A tetrode tube, operating at a frequency of 50 mc. The self-neutralizing frequency of the 4-125A is approximately 80 mc, so the tube may be neutralized in the normal manner, as shown in Figure 14. It is possible for this amplifier to develop a VHF parasitic oscillation in the

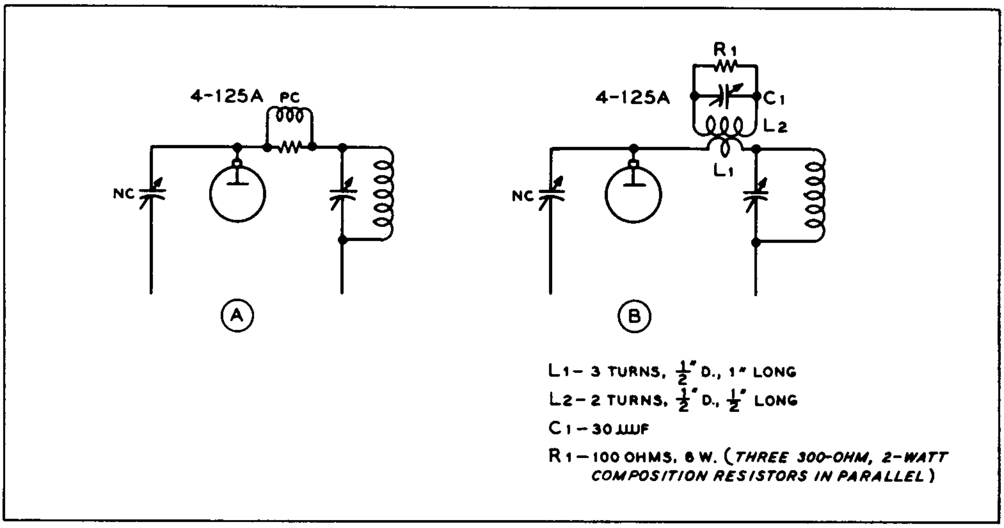


Fig. 15 Low frequency parasitic supression circuit (A) will absorb too much power at VHF. Inductively coupled, tuned trap should be used as substitute. Resonant circuit is loaded by noninductive resistor to provide circuit loss at parasitic frequency, yet allow high impedance at operating frequency.

region of 150 mc, well above the self-resonant frequency of the tube. Utilizing either of the two methods of VHF neutralization described in Figures 12 and 13 will tend to make the circuit unstable at 50 mc, and the neutralization of the stage at 50 mc will tend to enhance the 150 mc parasitic circuit.

The solution to this dilemma is to introduce enough circuit loss at the parasitic frequency to load the circuit past the point of oscillation. A simple parasitic choke as shown in Figure 15A will suppress the parasitic, but at the same time will absorb enough of the fundamental frequency energy to cause a large drop in circuit efficiency. The suppressor will probably burn up under these circumstances. If, however, the parasitic choke is tuned to the parasitic frequency and coupled to the plate circuit of the tube as shown in Figure 15B, a minimum of fundamental frequency energy will be dissipated in the choke. The amplifier circuit will then respond to neutralization at the operating frequency, without danger of parasitic oscillation at some frequency above the self-neutralizing frequency of the tube.

This circuit may also be employed with triode tubes operating in the VHF region if parasitic oscillations are encountered.

VHF Television Interference

Various forms of TVI can plague the VHF amateur as well as the low frequency operator. The causes of VHF TVI, in general, may be traced to the same basic causes as those which produce so much trouble in the low frequency portion of the spectrum.

The main cause of TVI that may be traced directly to the VHF transmitter is the generation and radiation of spurious harmonic and parasitic signals. The elimination of parasitic oscillations has been covered earlier in this chapter. By proper equipment design, the level of spurious harmonic radiation may be attenuated to such a value as to cause a minimum of interference to nearby television receivers.

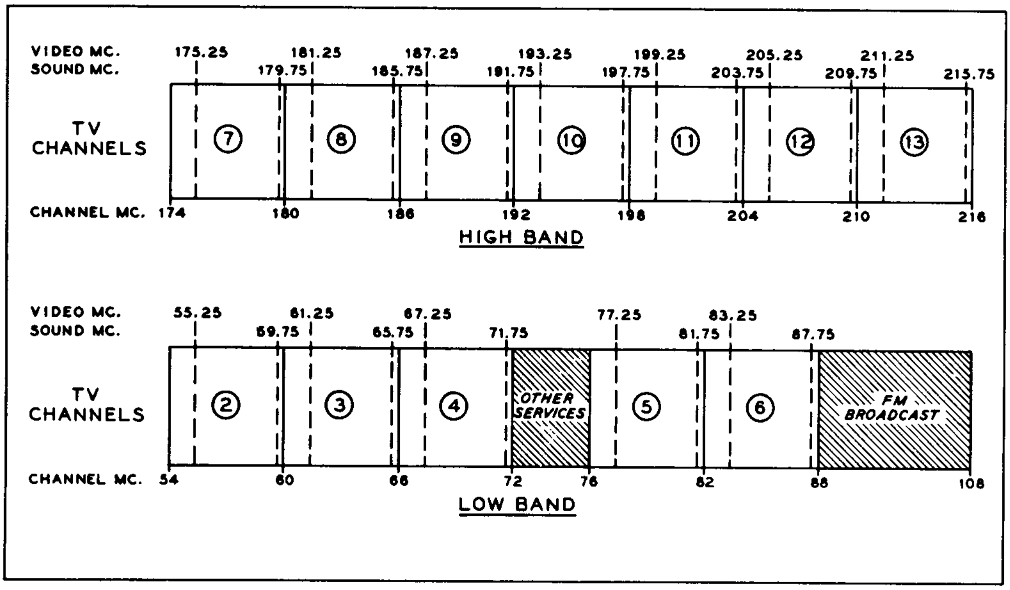


Fig. 16 VHF television channel chart, showing video and sound frequencies.

Harmonic Generation

A chart of the various VHF television channels is shown in Figure 16. It can be seen that the second harmonic of 50 mc stations will fall into the FM broadcast band (88 mc to 108 mc) causing possible interference in that region. The third harmonic of the six meter transmitter will fall within the range of TV channels 11, 12, and 13. The ninth and tenth harmonics fall within the limits of the UHF television channels. Harmonics of a two meter transmitter will also fall into the UHF channels.

Unwanted harmonics generated by the exciter stages of VHF transmitters can cause TVI in other channels than those listed above. In particular, when exciter stages are operated on 6, 8, 25, or 48 mc, severe harmonic interference to certain TV channels may develop as shown in Figure 16. If capacity coupling is employed between the doubler stages of an exciter, the harmonics of the fundamental frequency can pass directly through the transmitter and be radiated by the antenna system. Thus a 50 mc transmitter employing a 6.25 mc crystal can possibly radiate harmonics at 56.25 mc, 62.5 mc, 68.75 mc, and higher multiples of 6.25 mc. These harmonics could cause TVI in all the low frequency TV channels. This interference may be cured by the use of inductive coupling between the doubler stages of the transmitter, eliminating the circuit response to the spurious frequencies. In addition, use of a high frequency crystal to eliminate some of the doubler stages will be a great help. The use of a 25 mc crystal is recommended for 50 mc operation, as harmonics of the crystal oscillator will then occur at 25 mc intervals, with the third, fourth, fifth and sixth harmonics falling outside the television channels. The seventh harmonic of a 25 mc oscillator falls within TV channel 7, but the proper use of inductive coupling can effectively eliminate trouble from this source.

The same trouble can show up in 144 mc transmitters that employ 9 mc or 18 mc crystals with doubler strings driving the power amplifier. A

string of harmonics of 9 mc or 18 mc oscillator stages can raise havoc with nearby television receivers. Inductive coupling between stages will help to reduce this trouble.

Harmonic Radiation

Generation of unwanted harmonics can cause trouble when these harmonics are radiated to a nearby TV receiver. Therefore steps should be taken to reduce harmonic radiation to a minimum. It is possible to choose the operating frequency of the transmitter so that the radiated harmonics fall into an unused TV channel. For example, 8 mc crystals can cause tenth harmonic interference to channel 6 from a 50 mc transmitter. The use of 25 mc crystals would move the harmonic out of channel 6, dropping the seventh harmonic of 25 mc into channel 7.

For best harmonic suppression, it is necessary to employ shielding around the power generating circuits of the transmitter, and to filter all leads leaving the transmitter. A discussion of shielded enclosures earlier in this chapter covers the correct techniques, which apply equally to TVI suppression as well as to energy radiation.

The use of shielded wiring for all power leads within the shielded enclosure will minimize the pickup of stray r-f energy by these leads and simplify the problem of filtering the lead at the point it leaves the enclosure. The shield should be grounded at each end of the lead, and at convenient points along its length.

Other TVI Problems

A second cause of VHF TVI may be traced to shortcomings of the usual

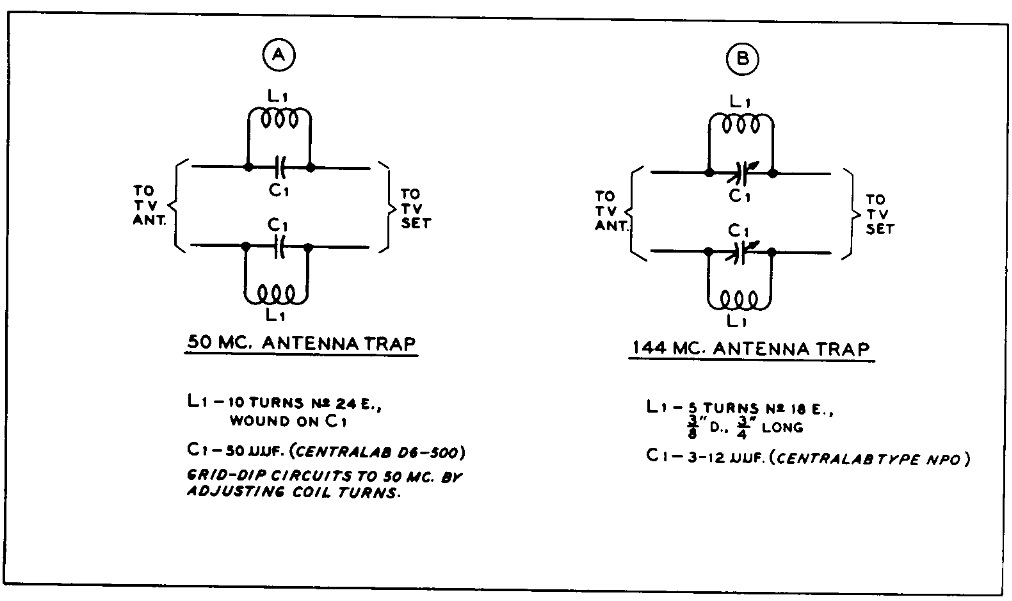
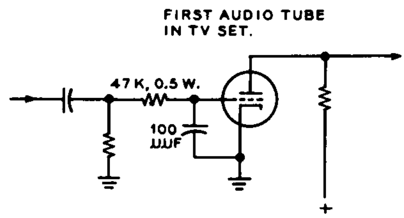


Fig. 17 Simple parallel resonant traps may be placed in each lead of TV receiver feedline to attenuate amateur signals in the 50 and 144 mc bands.

Fig. 18 R-C filter in grid lead of first audio stage of TV receiver will eliminate interference caused by audio rectification of signal.



television receiver. The TV set may not have enough adjacent channel selectivity to reject the signals of a nearby amateur station, particularly if the station is operating on the 50 mc band, and the TV set is tuned to channel 2. The amateur signal will overload various circuits of the receiver, causing blocking to occur. Blocking may be reduced by placing a tunable trap in the 300 ohm transmission line directly at the antenna terminals of the television receiver. A suitable trap for rejection of signals in the 50 mc region is shown in Figure 17A.

Channel 2 interference may be encountered on the 144 mc band on TV receivers having an intermediate frequency of 45 mc. The image signal of the amateur station will fall within the tuning range of channel 2 if the TV set does not have a sufficiently high order of image rejection. A 144 mc tuned trap placed at the antenna terminals of the television set, as shown in Figure 17B will help to reduce this type of interference.

Still another type of TVI may be encountered in VHF operation wherein the audio portion of the TV receiver exhibits some degree of nonlinearity and acts as a detector, rectifying the audio portion of the amateur signal and superimposing it upon the TV sound signal. A simple audio filter placed in the television set will usually cure this difficulty (Figure 18).

The use of suitable filter circuits in both the amateur transmitter and the nearby television receiver can do much to eliminate these forms of interference.

Low Pass TVI Filters for the VHF Range

Three low pass filters for VHF transmitters are described herewith. The first filter may be used with any transmitter operating below 54 mc. The filter provides protection for television channels 3 through 13, and the UHF channels. A filter of this type cannot be used to protect channel 2, as the cutoff characteristic of the filter is not sharp enough. Channel 2 protection can only be obtained by a very elaborate filter having low insertion loss in the 50 mc band. Construction of such a filter is beyond the scope of this book.

The 2 meter low-pass filter shown in Figure 19 is built in an interlocking aluminum box measuring 1⅝" x 2⅛" x 6½" *(L.M.B. #650)*. The box is divided into three similar compartments by the insertion of two partitions cut from thin brass shim stock, obtainable at any large hardware store.

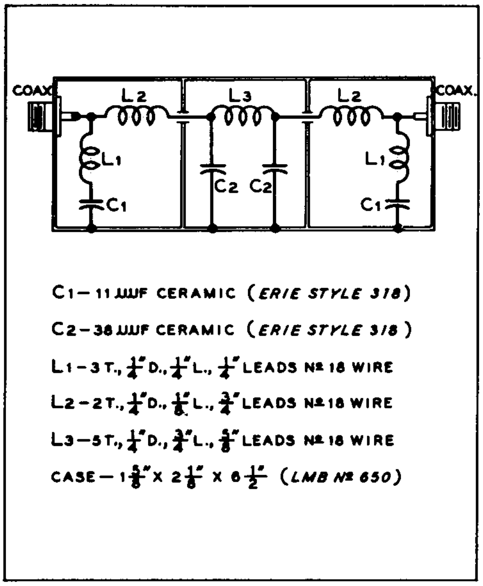


Fig. 19 This two meter low pass filter provides maximum value of rejection above 170 mc. The filter is designed for 52 ohm coaxial line, but will perform well in 72 ohm coaxial lines. Filter is constructed within a three section aluminum box.

These partitions are bolted in place by means of 4-40 machine screws and bolts. Coils L1 and L2 should be mounted at right angles to one another to reduce mutual coupling to a minimum. Capacitors C2 are used to support the junctions of coils L2 and L3. If a grid dip oscillator is at hand, the end sections of the filter may be checked by shorting the coaxial plugs at the junction of L1 and L2 with a short, heavy strap and adjusting the turns of L1 so that circuit L1-C1 is resonated to 200 mc. The next step is to temporarily disconnect the two L2 coils from C2 and L3, and grid dip the circuit C2-L3-C2 to 112 mc, by squeezing the turns of L3. The coils should now be resoldered in position, and the filter is ready for use.

The second low-pass filter shown in Figure 20 is a version of the *General Electric "Harmoniker."* It is a half-wave filter with the attenuation increasing 30 db per octave above the operating frequency. Second harmonic

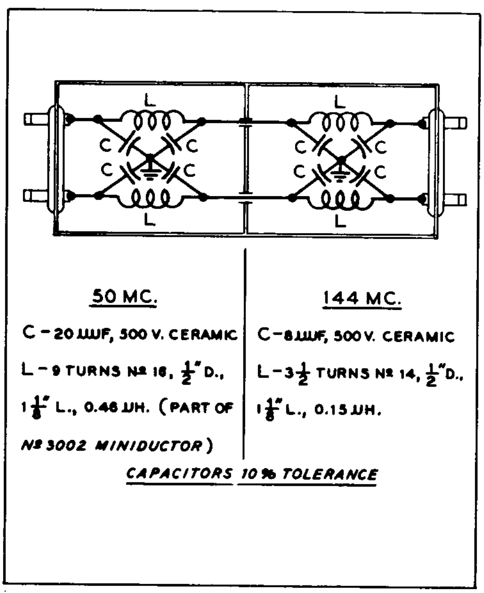


Fig. 20 The "Harmoniker" type filter may be used on the 50 mc and 144 mc bands. This filter is designed for use with 300 ohm ribbon or open wire lines. The case of the filter should be grounded to the transmitter.

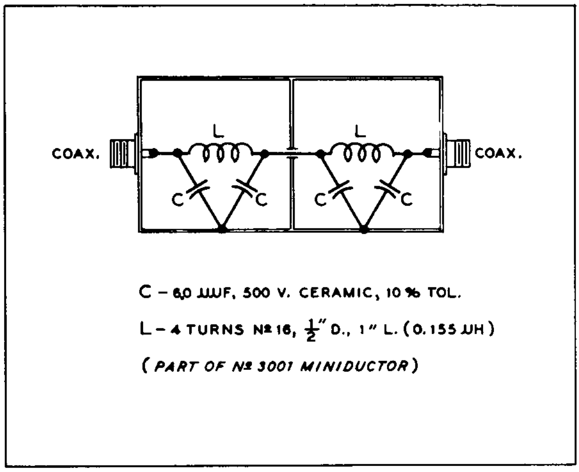


Fig. 21 "Harmoniker" type filter may be divided in half for use with coaxial line. Filter at right is intended for use in two meter band with 52 or 72 ohm coaxial lines.

attenuation is thus 30 db, third harmonic—48 db, fourth harmonic—59 db, and so on. Coil information for 50 mc and 144 mc filters is provided in the drawing. The filter is intended for use with balanced transmission lines having impedances of 300 to 500 ohms. The unit may be built in an aluminum interlocking box measuring 3″ x 4″ x 5″ *(L.M.B. #140)*. A single brass partition is placed across the center of the box, and two ½″ holes are drilled in the partition to pass the coil leads. *Millen* 37412 plugs and 33012 sockets may be used for terminal connectors.

The third filter is a single-ended Harmoniker unit, intended for use with coaxial transmission lines on the 144 mc band. The unit is constructed in an interlocking aluminum box measuring 4″ x 2⅛″ x 1⅝″ *(L.M.B. #00)*. A single brass partition is placed across the center of the box, with a ½″ hole drilled in the center to pass the lead connecting the two coils. The capacitors are grounded to common points within each compartment with the shortest possible leads. Upon completion, the unit may be checked by soldering a low inductance ground strap to the lead passing through the hole in the center partition. The coils are adjusted so that each section of the filter resonates to 145 mc. The ground strap is then removed, and the filter is ready for use.

VHF Assemblies

Because of the fact that all components in a VHF circuit have inherent properties that do not aid the operation of the circuit, it is necessary that the minimum number of the smallest size parts be employed in VHF assemblies. The layout of the principal parts should be arranged so that the components themselves provide the necessary lead connections to each other, without the use of long, interconnecting wires. When the components are small, a great many of them may be mounted directly upon the socket pins of the r-f tubes. Other components may be mounted between socket pins and adjacent bypass condensers and tie points.

CHAPTER VI

VHF Antenna Design

The role of the antenna in a radio communication circuit is to serve as a coupling device which converts electronic energy supplied by the transmitter to electrostatic and electromagnetic waves which are propagated through the ether. At the receiving station a similar antenna system converts these emanations back to electronic energy which can be detected and demodulated by the receiving equipment. There are many antenna configurations, and a complete study of these would cover many volumes. Within the space limitations of this book, the antenna parameters most important to VHF circuits will be discussed, and antenna designs which have proven to be practical in the amateur bands will be illustrated and analyzed.

Antenna Terminology

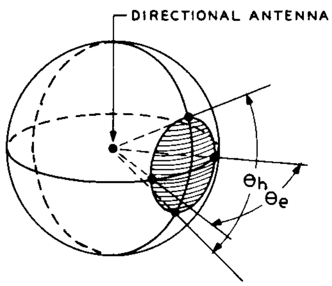
Certain terms and characteristics peculiar to antenna systems should be defined as the starting point of this discussion.

Directivity

In the case of a transmitting antenna, directivity is defined as the ability of the antenna to concentrate radiation in a particular direction. All practical antennas exhibit some degree of directivity. A completely nondirectional antenna (one which radiates equally well in all directions) is known as an *isotropic radiator,* and only exists as a mathematical concept. Such a radiator if placed at the center of an imaginary sphere would "illuminate" the inner surface of the sphere uniformly.

When speaking of a receiving antenna, directivity will be always reciprocal to the transmitting characteristics of the antenna. That is, the antenna will intercept waves best from the same directions in which it concentrates energy when employed as a radiator. Thus, an antenna providing a given set of radiation requirements will provide equal results when used for receiving.

FIGURE 1



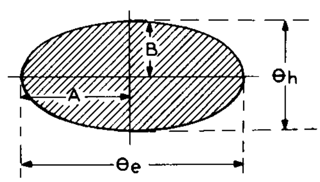
$$\text{POWER GAIN OVER ISOTROPIC RADIATOR} = \frac{\text{SURFACE AREA OF SPHERE}}{\text{AREA OF ELLIPSE AT HALF-POWER ANGLES}} \qquad (1)$$

ONE RADIAN = 57.324 DEGREES

THE AREA OF A SPHERE IS EQUAL TO: 4π SQUARE RADIANS (2)

THE AREA OF AN ELLIPSE (OR A CIRCLE) IS EQUAL TO: πAB SQUARE RADIANS (3)

WHERE A *AND* B *ARE ONE-HALF THE LENGTH AND WIDTH, RESPECTIVELY, OF THE ELLIPSE EXPRESSED IN RADIANS.*



θ_e *AND* θ_h *REPRESENT THE HALF-POWER BEAM WIDTHS IN THE ELECTRIC AND MAGNETIC PLANES, RESPECTIVELY. THE ELECTRIC PLANE IS GENERATED IN THE SAME PLANE AS THE RADIATOR ELEMENT, WHILE* θ_h *IS GENERATED IN THE PERPENDICULAR PLANE.*

$$A \text{ IN RADIANS} = \frac{\theta_e}{114.59} \qquad B \text{ IN RADIANS} = \frac{\theta_h}{114.59} \qquad (4)$$

$$\text{THEREFORE: } G = \frac{4\pi}{\pi \frac{\theta_e\,\theta_h}{(114.59)^2}} = \frac{52525}{\theta_e\,\theta_h} = \text{POWER GAIN OVER ISOTROPIC RADIATOR} \qquad (5)$$

SINCE A HALF-WAVE DIPOLE HAS A GAIN OF 1.64 OVER AN ISOTROPIC RADIATOR, THE GAIN OF A DIRECTIONAL ANTENNA OVER A HALF-WAVE DIPOLE MAY BE EXPRESSED AS:

$$\text{POWER GAIN (G)} = \frac{52525}{(1.64)\theta_e\,\theta_h} = \frac{32027}{\theta_e\,\theta_h} \qquad (6)$$

AN ILLUSTRATION OF HOW POWER GAIN IS THE RATIO BETWEEN THE SURFACE AREA OF A SPHERE ILLUMINATED BY AN ISOTROPIC RADIATOR AND THE PORTION OF THE SPHERE WHICH LIES BETWEEN THE HALF-POWER ANGLES, θ_e AND θ_h, OF THE DIRECTIONAL RADIATOR.

Power Gain

Power gain in an antenna system is a term used to express the power increase in the radiated field of one antenna as compared to a standard antenna. The power gain is the product of directivity and radiation efficiency, and is taken to be measured in the direction of maximum field energy. For example, a half wave dipole produces 1.64 times the power at a distant point than that produced by an isotropic radiator under similar conditions. This is equivalent to a power gain of 2.1 decibels.

Since an isotropic radiator exists only in theory, it is common to refer power gain to a half wave dipole. In other words, if a certain directional antenna delivers ten times the power in its favored direction as compared to a half wave dipole, the directional array has a gain of 10 decibels (db) over the reference dipole. Thus the expression of power gain of any antenna means nothing unless the reference source is defined. This reference is often omitted in published gain figures of manufactured antennas. In this Handbook, all power gain figures are referred to a standard dipole, and not to an isotropic radiator. The physical concept of power gain is illustrated in Figure 1, showing the illumination of a sphere by a directional antenna.

Calculation of Power Gain

By definition, an isotropic radiator would illuminate the sphere of Figure 1 in a uniform manner. The power gain of any antenna placed at the center of the sphere may be thought of as the ratio between the surface area of the sphere illuminated by the isotropic radiator and the portion of the sphere illuminated by the directional antenna. Since the field pattern of any directional antenna is not clear cut but blends into nothingness at the extremities, the actual pattern is defined as that elliptical, illuminated portion of the sphere which lies between the "half-power" angles of the radiator field. On the usual polar plot of a beam pattern, these points are the "—3 db" power points.

The power gain of any antenna over a dipole may be computed from Figure 1 when the beam width (expressed in degrees) in the cross section of the plane of radiation is known. Formula (6) is one well worth remembering, as it provides a quick method for determining power gain over a dipole antenna.

The preceding calculations have assumed that no power is spent in backward radiation or in spurious lobes that may actually exist in a directional antenna. If the spurious lobes are 15 db or more below the strength of the main lobe, the formulas are quite accurate. Even if side lobes are down only 10 db, the calculation of gain will be correct within two- or three-tenths of a decibel.

It should be noted that directivity alone does not always provide a true picture of power gain. Radiation efficiency may at times be an important factor, since an antenna can have excellent directional characteristics but at the same time be crippled by poor radiation efficiency. If much of the transmitter power is absorbed and converted to heat in the antenna structure, the actual power gain may be considerably less than the directivity pattern would indicate.

From the receiving standpoint, the power gain of the antenna will be the same as when it is transmitting, and will manifest itself as an increase

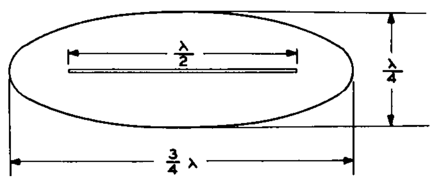
FIGURE 2

IF THE POWER GAIN OF AN ANTENNA SYSTEM IS KNOWN, THE EFFECTIVE APERTURE MAY BE CALCULATED FROM:

$$A_{eM} = \frac{1.64\,G}{4\pi} = 0.13\,G' \qquad (7)$$

WHERE G′ *IS THE POWER GAIN OVER A HALF-DIPOLE.*

FROM (7), THE EFFECTIVE APERTURE OF A HALF-WAVE DIPOLE IS EQUAL TO 0.13 SQUARE WAVELENGTHS. THE APERTURE IS ELLIPTICAL, AND MEASURES APPROXIMATELY 3/4 X 1/4 WAVELENGTH.



IT MAY BE MORE CONVENIENT TO FIND THE EFFECTIVE APERTURE IN SQUARE FEET, IN WHICH CASE THE FORMULA BECOMES:

$$A_{eM}\ (\textit{SQ. FT}) = \frac{G f^2}{7{,}586{,}000} \qquad (8)$$

WHERE G *IS THE GAIN OVER A HALF-WAVE DIPOLE AND* f *IS THE FREQUENCY IN MEGACYCLES.*

THE FOLLOWING FORMULAE MAY BE USED TO CALCULATE THE ACTUAL DIMENSIONS OF THE APERTURE:

--WHEN THE APERTURE IS CIRCULAR (OR ELLIPTICAL), SUCH AS WOULD BE THE CASE WHEN USING 2 STACKED YAGIS:

$$A_h = 2\sqrt{\frac{A_{eM}\,\theta_h}{\pi\,\theta_e}} \qquad A_e = 2\sqrt{\frac{A_{eM}\,\theta_e}{\pi\,\theta_h}} \qquad (9)$$

WHERE A_h *IS THE APERTURE WIDTH IN THE "h" PLANE, AND* A_e *IS THE APERTURE WIDTH IN THE "e" PLANE.*
A_h *AND* A_e *ARE EXPRESSED IN WAVELENGTHS.*

--WHEN THE APERTURE IS RECTANGULAR (OR SQUARE), SUCH AS WOULD BE THE CASE WHEN USING MANY PHASED DIPOLES:

$$A_h = \sqrt{\frac{A_{eM}\,\theta_h}{\theta_e}} \qquad A_e = \sqrt{\frac{A_{eM}\,\theta_e}{\theta_h}} \qquad (10)$$

--TO CONVERT A_h, *(OR* A_e*) FROM WAVELENGTHS TO FEET:*

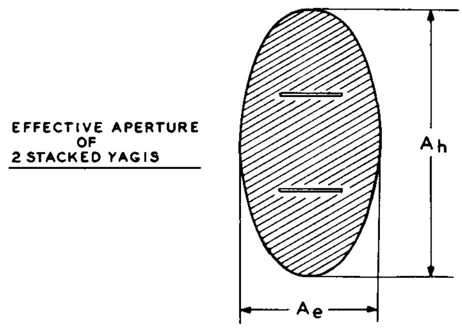
$$A_h\ \textit{IN FEET} = \frac{A_h \text{ IN WAVELENGTHS} \times 984}{\text{FREQUENCY IN MEGACYCLES}} \qquad (11)$$

DETERMINATION OF EFFECTIVE APERTURE OF AN ANTENNA

FIGURE 3

EXAMPLE: *CALCULATE THE PHYSICAL SIZE OF THE ELLIPTICAL APERTURE GENERATED BY A 2-YAGI ARRAY.*

HALF-POWER BEAM WIDTH IN THE "e" PLANE IS 20 DEGREES.
HALF-POWER BEAM WIDTH IN THE "h" PLANE IS 10 DEGREES.



THE POWER GAIN OVER A HALF-WAVE DIPOLE, $G = \frac{32027}{10\,(20)} = 160.1$ (22.05 DB)

EFFECTIVE APERTURE, $A_{eM} = 0.13\,(160.1) = 20.8\,\lambda^2$

SINCE THE APERTURE IS ELLIPTICAL, WE WILL USE THE FORMULA (9)

$$A_e = 2\sqrt{\frac{20.8\,(10)}{\pi\,(20)}} = 3.62 \text{ WAVELENGTHS}$$

$$A_h = 2\sqrt{\frac{20.8\,(20)}{\pi\,(10)}} = 7.26 \text{ WAVELENGTHS}$$

AT 144 MEGACYCLES, A_e *IN FEET* $= \frac{3.62\,(984)}{144} = 24.7$ FEET

A_h *IN FEET* $= \frac{7.26\,(984)}{144} = 49.6$ FEET

THE NOMOGRAPH OF FIG. 5 GIVES ANTENNA GAIN AND APERTURE AREA DIRECTLY FROM "e" AND "h" PLANE BEAM WIDTHS, AND WILL PROVE HELPFUL IN RAPID CALCULATIONS.

EXAMPLE OF GAIN AND APERTURE CALCULATION OF YAGI ARRAY

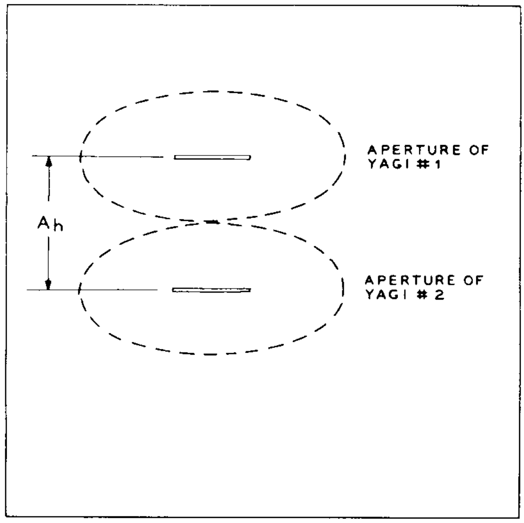


Fig. 4 Antennas in a stack may be spaced so that apertures just "touch." This will allow maximum gain consistent with minimum stacking distance. High gain arrays should be stacked several wavelengths apart to prevent overlap of apertures, as shown at right.

in the amount of signal power delivered to the transmission line as compared to that amount produced by a reference antenna.

Effective Aperture

Effective aperture (Aem) is closely associated with directivity and power gain. In a simplified analogy it may be thought of as the frontal area from which the receiving antenna will extract signal power from the radio wave. Sometimes this concept is referred to as *capture area.* On the other hand, *physical aperture* is a measure of the area physically occupied by the antenna system, and may be smaller or greater than the effective aperture. Some antennas, such as the parabolic dish type have an effective aperture somewhat smaller than their physical size, while others such as high-Q parasitic arrays have an effective aperture considerably larger than their physical size. The ratio between the effective and the physical apertures is known as the *K-factor,* and indicates what order of power gain can be expected from an antenna of given size. The higher the K-factor, the higher will be the gain per unit size of the antenna, although other factors such as bandwidth and spurious lobes must be taken into consideration when choosing an antenna.

Spacing Between Arrays as Determined by Apertures

One of the important reasons for determining effective aperture size is that it provides us with a key as to the distance we should use to space antennas when stacking two or more in an array. A simple rule of thumb is to make the spacing such that the effective apertures just "touch" one another as shown in Figure 4. If the antennas are spaced closer than this, the effective apertures will overlap, and power gain will be reduced. The higher the gain figure of the individual antennas, the larger will be the effective aperture, thus calling for increased spacing between bays. Dipoles may be spaced one-half wavelength in an array without overlap of apertures, whereas high gain Yagis used in an array must be spaced several wavelengths apart to prevent overlap of apertures.

Gain and Aperture Nomograph

The relationships between power gain, effective aperture and beam widths may be expressed in a simple nomograph, as shown in Figure 5. Antenna gain and aperture area may be derived from E and H plane beam widths. The graph assumes that all spurious antenna lobes are down 10 db or more below the strength of the main lobe. As shown in the example, an antenna array having a beam width of 10 degrees by 20 degrees at the half power points will produce a power gain of 22 db, and will have an effective aperture of 21 square wavelengths. The dimensions of the aperture may be calculated from formulas (9) and (10).

The Wavelength Factor

One of the most interesting, yet least understood aspects of the VHF communication circuit is the relationship between operating frequency and the physical size of the antenna. At first glance it might seem logical to assume that it is possible to reduce antenna dimensions in direct relationship to frequency and end up with similar results. In other words, it does not seem illogical to assume that half wave dipoles at each end of a 144 mc circuit will perform as well as half wave dipoles at each end of a 50 mc circuit. Unfortunately, the assumption is not true, as the wavelength factor enters the picture. The 50 mc dipole is roughly three times as long as the 144 mc dipole, and will therefore extract about three times as much energy from the radio wave as will the smaller antenna. Even though the relation between antenna size and wavelength remains constant, the larger dipole will extract more energy from the passing wave than will the smaller dipole. On the other hand, during periods of transmission, the dipoles are equal as to radiation efficiency, and regardless of the wavelength will exhibit the same degree of directivity. Thus the wavelength factor is more important to the receiving system than to the transmitting system.

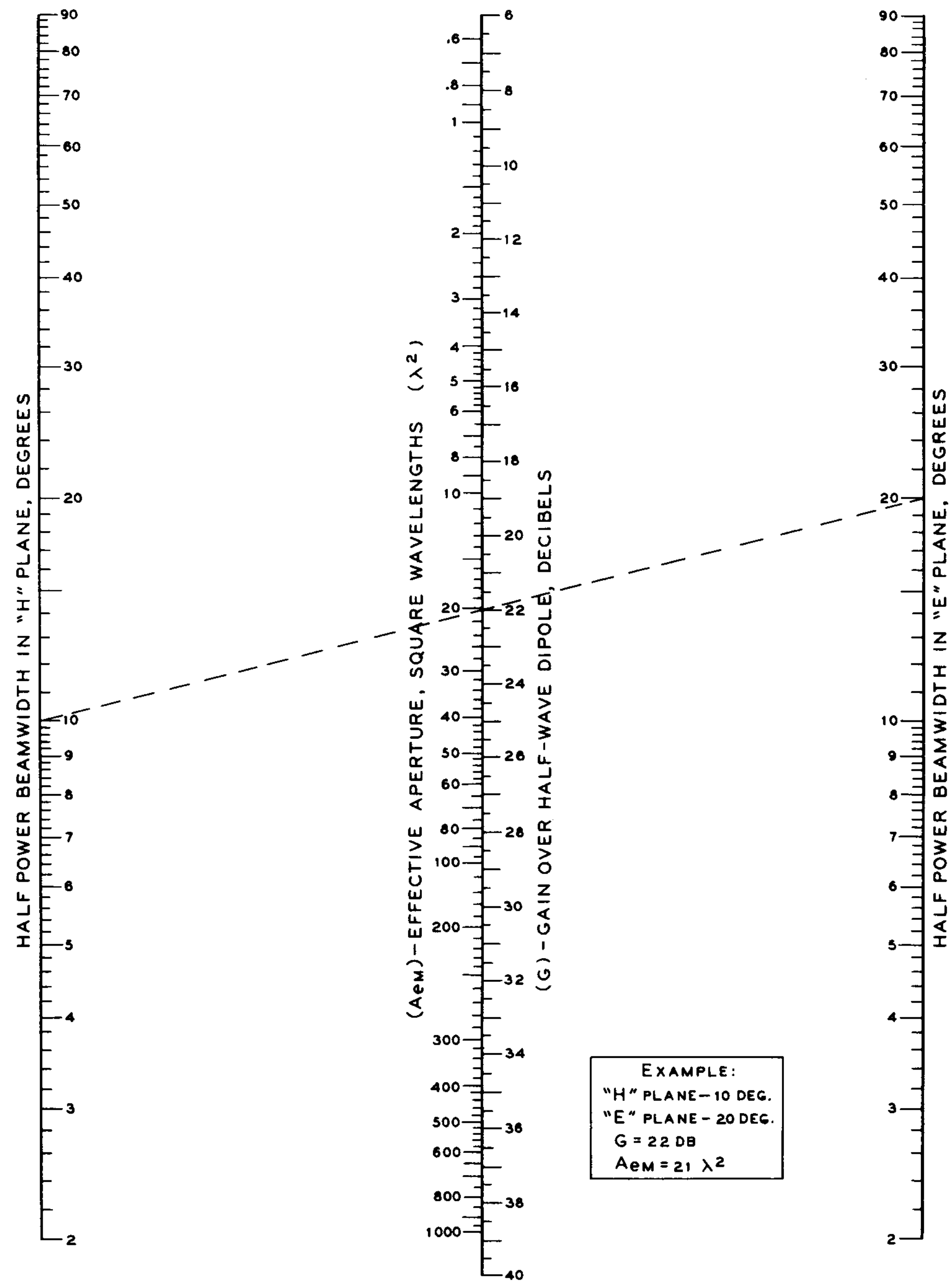
From this it can be seen that it is necessary to keep antenna size constant at the receiver, regardless of frequency, if we want overall circuit performance to remain the same. However, if antenna size is held constant as the operating frequency is increased, it follows that the directivity of the antenna must increase. Formula (7) shows that power gain (G) of an antenna increases as the aperture size is increased. As mentioned before, power gain and directivity are synonymous.

As a result of these conditions, choice of the optimum operating frequency becomes important. If antenna size is held constant, and we wish to achieve wide coverage with little directional effects from the antenna the lowest possible frequency should be employed. On the other hand, for point to point communication, especially when it is necessary to discriminate against interference, the highest frequency will be most effective. The higher frequency will provide just as much circuit efficiency provided the antennas are held to constant size, but greater directivity must then be accepted.

Polarization

With most forms of propagation used in the VHF region it is most im-

FIGURE 5



NOMOGRAPH FOR DETERMING ANTENNA GAIN AND APERTURE SIZE FROM BEAM WIDTHS IN "E" AND "H" PLANES.

NOTE: *THE GRAPH ASSUMES THAT THE ANTENNA HAS NO SPURIOUS LOBES. IF SPURIOUS LOBES ARE 10 DB OR MORE BELOW THE STRENGTH OF THE MAIN LOBE, THE GRAPH IS QUITE ACCURATE.*

portant that wave polarization be the same at each end of the circuit. This means that the active elements of the receiving and transmitting antennas must lie in the same relative plane. The two most commonly used planes of polarization are the horizontal and vertical. Only during the relatively rare instances of ionospheric skip is there enough depolarization of the radio wave to make antenna orientation unimportant.

Generally speaking, horizontal polarization (antenna elements parallel to the surface of the ground) holds a slight performance edge over vertical polarization on long scatter paths. It is also more effective in discriminating against local noise pickup, since most man-made electrical noise seems to be vertically polarized. For these reasons, horizontal polarization has gained prominent use in long distance circuits where every advantage must be taken. Vertical polarization is used extensively in local, short-distance communications where universal coverage is desired. Small, non-directional antennas are easier to achieve with vertical elements than with horizontal, especially in the case of mobile installations. A quarter wavelength whip, with the metal car top serving as a ground plane is a very popular VHF mobile antenna.

For the amateur bands, the best rule is for the "DX-hound" to employ horizontal polarization, while those amateurs interested more in local contacts can profitably use vertical polarization.

Antenna Height

The most important requirement of a good VHF antenna is that it be situated high enough to clear all nearby obstacles by at least two to three wavelengths. A high gain antenna lost in a forest of power lines, buildings, or trees may not be as effective as a simple dipole high above the ground. Once the antenna is high enough to be in the clear, the benefits of further height increase will depend upon the general terrain within a few miles of the station, and the type of antenna to be used. In general, greater antenna height will increase the field strength of ground wave propagation up to 100 miles or so, while on longer paths where tropospheric scatter becomes a major factor, antenna height is less important. Antenna height is not particularly important for propagation of ionospheric scatter, meteor reflection, sporadic-E, or F_2 skip transmission modes. Probably the most effective antenna combination for the VHF station is a large antenna array at a medium height, and a small array erected as high as possible. Each will exhibit distinct advantages under certain types of propagation.

Antenna Measurements

It has been quite a common practice for antenna adjustments to be made with the aid of a field strength meter located at some remote point. Such an arrangement can prove quite helpful when used properly, but any attempts to measure antenna gain, or to compare two antennas by measuring field strength along a ground path are subject to extreme inaccuracies. Factors such as ground reflection and the proximity effects of nearby objects obliterate reference factors and reduce the results to meaningless figures. Unless the test antenna and field strength meter are located many, many wavelengths above ground, the effect of ground reflection will throw doubt on each and every measurement. This condition cannot be emphasized too strongly. Many

published measurements have been in error due to what is apparently a lack of understanding of the pitfalls produced by ground reflection.

The most reliable method for measuring antenna performance is that of plotting the radiation pattern of the antenna. This is done by rotating the test antenna through 360 degrees and recording the change in relative field strength at a point 20 or more wavelengths away. By running this test in both the E and H planes of the antenna, it is possible to obtain a three dimensional view of the radiation pattern of the antenna which includes the spurious lobes and back radiation. Using this information the actual power gain of the antenna may be calculated with the aid of formulas (5) and (6). Only the half-power beam widths in the E and H plane are required for this experiment.

It is important that a calibrated field strength meter be used, since an accurate indication of the half power points of the beam pattern must be made. A good signal generator with an accurate attenuator can be used for calibrating the field strength meter. A typical set-up for measurements of this type is illustrated in Figure 6.

VHF Antenna Types

All forms of VHF antennas may be classified into three characteristic groups: *colinear*, *broadside*, and *end-fire* arrays. In addition, a large array may be composed of any two of these groups combined into a single entity, such as a colinear, end-fire array. It is suggested that the reader familiarize himself with antenna and transmission line fundamentals from such sources as the *Beam Antenna Handbook*, published by Radio Publications, Inc., Danbury Road, Wilton, Conn., the *Antenna Manual*, published by Editors and Engineers, Summerland, Calif., or the *A.R.R.L. Antenna Book*, published by the American Radio Relay League, West Hartford, Conn.

Colinear Arrays

A colinear array is one in which the elements lie in the same line end to end, and are fed with equal in-phase energy, as shown in Figure 7. The

Fig. 6 Field strength meter should be located several wavelengths above antenna under test to minimize effects of ground reflection. Radiation pattern of array is checked to observe included angle at half-power points of pattern.

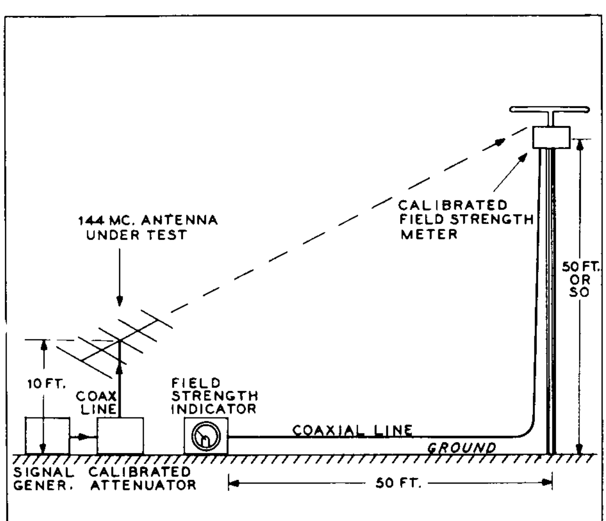
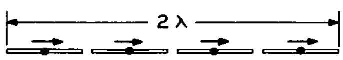
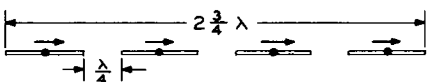


FIGURE 7

COLINEAR ARRAY

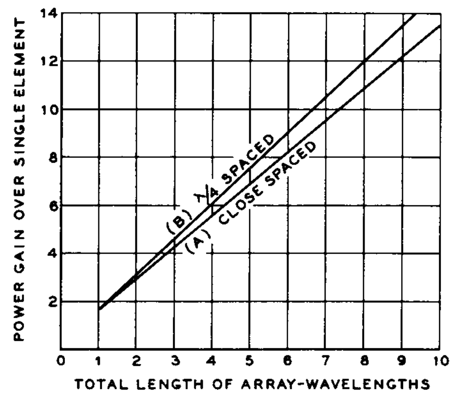


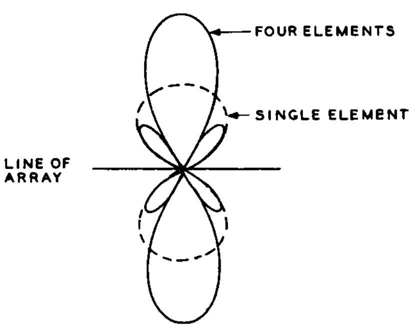
(A) CLOSE SPACED



(B) WIDE SPACED

NOTE THAT RADIATION IS PERPENDICULAR TO THE PAGE.





(C) TYPICAL RADIATION PATTERN

ALL ELEMENTS ARE DRIVEN WITH EQUAL CURRENTS IN PHASE. THE WIDER SPACING GIVES HIGHER GAIN BUT IS MORE DIFFICULT TO FEED. THE CLOSE SPACED ELEMENTS CAN BE COUPLED WITH QUARTER-WAVE STUBS, OR HALF-WAVE PHASING SECTIONS FOR CONVENIENT FEEDING. A SET OF PARASITIC REFLECTORS, OR A REFLECTOR GRID PLACED A QUARTER WAVELENGTH BEHIND WILL PRODUCE A UNIDIRECTIONAL PATTERN AND AN INCREASE OF ABOUT 3 DB IN POWER GAIN.

separation between the tips of the elements is not critical, although one-quarter wavelength spacing between the half wave elements produces a slightly enhanced gain figure at the expense of the physical size of the array and feed point complications. Maximum radiation is at right angles to the line of the array, as shown in Figure 7C.

Broadside Arrays

A broadside array is one in which all elements are fed with equal in-phase energy, and lie in the same plane. The radiation pattern is perpendicular to the axis of the array and the plane containing the elements. Spacing between the broadside elements is of the order of one-half to one wavelength. Minor spurious lobes appear in the beam pattern as the element spacing is increased above one-half wavelength. The configuration and power gain of typical broadside arrays are shown in Figure 8.

End-Fire Arrays

An end-fire array is one in which the direction of radiation coincides with the direction of the axis of the array. The elements may be fed equal currents with progressive phase shift, or (in the more usual case) they may be parasitically excited from one fed element. In this case, the end-fire array is termed a *Yagi antenna.* The parasitic elements obtain their excitation solely by proximity coupling between themselves and the fed element. The parasitic element closest to the fed element lying in the line of fire of the array is termed the *launching element.* The parasitic-type Yagi antenna provides more power gain per unit of cost and per unit of space and weight than any other type of array. The configuration and power gain characteristic of Yagi and all-driven end-fire arrays are illustrated in Figure 9.

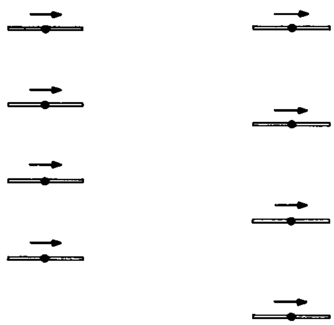
Array of Arrays

It is common practice in VHF work to combine various types of arrays to form large directional antennas, exhibiting high gain figures and very narrow beam widths. The chart of Figure 10 indicates the power gain which may be produced by various combinations of colinear and broadside radiators, backed up by parasitic reflector elements. It is important that the radiating elements be fed with equal currents, and in phase with one another. This becomes increasingly difficult as the number of elements excited from one feedpoint increases. The H-array of Figure 10A is an effective method of driving four radiators. Careful dimensioning will produce good results in feeding eight radiators, as shown in Figure 10B. Beyond this number, it is best to break the array up into groups of four or eight radiators as in (A) and (B) respectively, feeding the groups as shown in Figure 10C. Two or more such sub-groups may be fed from a manifold-type harness to achieve the power gains listed in the chart. By allowing one-quarter wave colinear spacing between groups, the power gain will be slightly higher than the value indicated in the chart.

If a reflector grid is used instead of parasitic reflectors, greater front to back ratio will be achieved, and the operational bandwidth of the array will be somewhat broader. Forward gain will, however, be only slightly higher.

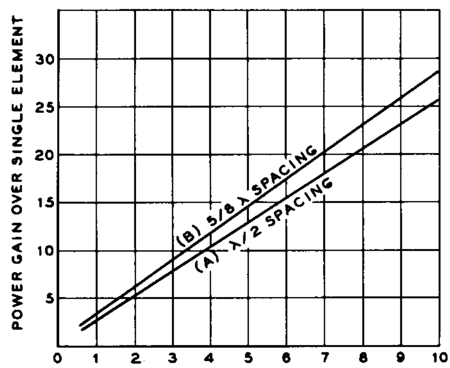
FIGURE 8

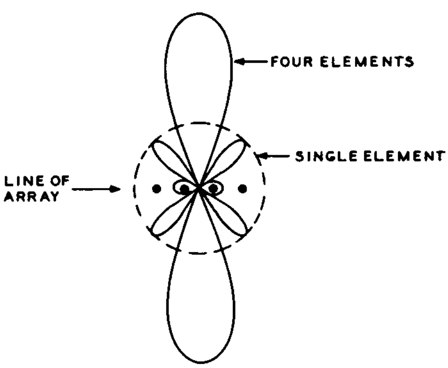
BROADSIDE ARRAY



(A) λ/2 SPACING (B) 5/8 λ SPACING

NOTE THAT RADIATION IS PERPENDICULAR TO THE PAGE.



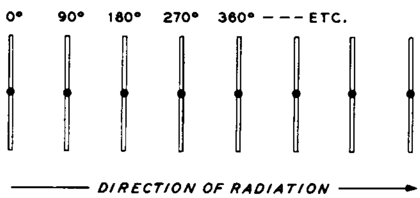


(C) TYPICAL RADIATION PATTERN

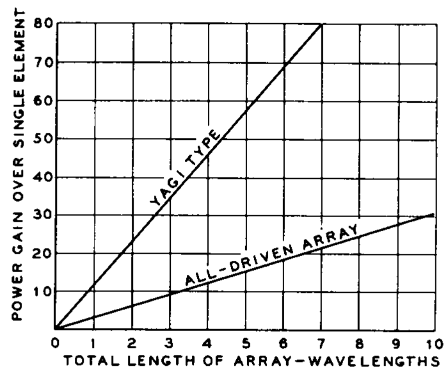
ALL ELEMENTS ARE DRIVEN WITH EQUAL CURRENTS IN PHASE. THE 5/8 WAVE SPACING GIVES HIGHER GAIN, BUT 1/2 WAVE IS MORE COMMON SINCE THE FEED SYSTEM IS SOMEWHAT SIMPLER. A SET OF PARASITIC REFLECTORS, OR A REFLECTOR GRID PLACED A QUARTER WAVELENGTH BEHIND WILL PRODUCE A UNIDIRECTIONAL PATTERN AND AN INCREASE OF ABOUT 3 DB IN POWER GAIN.

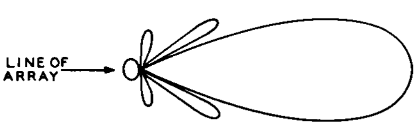
FIGURE 9

END-FIRE ARRAY



(A) ELEMENTS ARE SPACED 1/4 λ APART





(B) TYPICAL RADIATION PATTERN

THE ELEMENTS MAY BE FED EQUAL CURRENTS WITH PROGRESSIVE PHASE SHIFT CORRESPONDING TO THE SPACING. OR ONLY THE SECOND ELEMENT NEED BE FED, THE REST BECOMING SHOCK EXCITED, PARASITIC ELEMENTS; IN OTHER WORDS, A YAGI TYPE. THE PROBLEMS OF FEEDING ALL THE ELEMENTS IN AN END-FIRE ARE QUITE COMPLEX. THE PARASITIC TYPE IS EASIER TO FEED AND PROVIDES HIGHER GAIN, BUT ELEMENT LENGTHS ARE QUITE CRITICAL AND BANDWIDTH IS RELATIVELY NARROW.

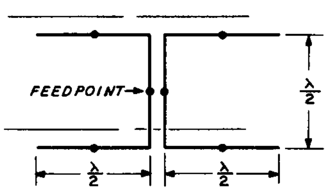
IN APPLICATIONS WHERE BANDWIDTH IS NOT TOO IMPORTANT, THIS TYPE WILL PROVIDE HIGHER GAIN PER CUBIC FOOT AND POUND OF ANTENNA THAN ANY OTHER TYPE.

FIGURE 10

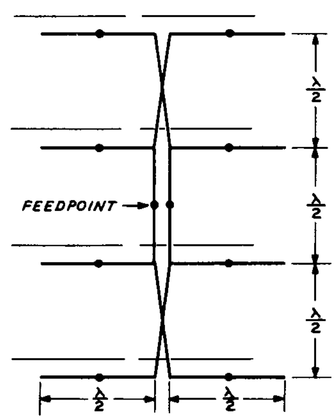
COLINEAR-BROADSIDE COMBINATIONS, WITH REFLECTORS

NUMBER OF COLINEAR ELEMENTS

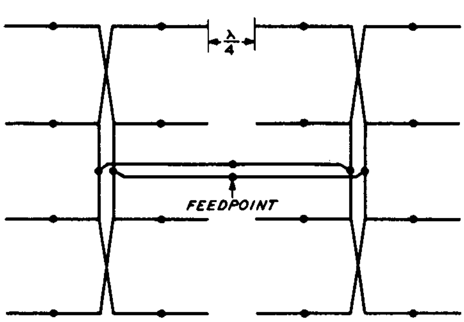
NUMBER OF BROADSIDE ELEMENTS	1	2	3	4	5	6	7	8
1	3.0	4.8	6.4	7.6	8.6	9.5	10.3	10.9
2	7.0	8.8	10.4	11.6	12.6	13.5	14.3	14.9
3	8.5	10.3	11.9	13.1	14.1	15.0	15.8	16.4
4	9.8	11.6	13.2	14.4	15.4	16.3	17.1	17.7
5	10.8	12.6	14.2	15.4	16.4	17.3	18.1	18.7
6	11.5	13.3	14.9	16.1	17.1	18.0	18.8	19.4
7	12.2	14.0	15.6	16.8	17.8	18.7	19.5	20.1
8	12.7	14.5	16.1	17.3	18.3	19.2	20.0	20.6
9	13.2	15.0	16.6	17.8	18.8	19.7	20.5	21.1
10	13.7	15.5	17.1	18.3	19.3	20.2	21.0	21.6
11	14.1	15.9	17.5	18.7	19.7	20.6	21.4	22.0
12	14.5	16.3	17.9	19.1	20.1	21.0	21.8	22.4



(A) 8 ELEMENTS.
FEEDPOINT IMPEDANCE IS AROUND 600-800 OHMS, DEPENDING ON CONDUCTOR SIZE AND SPACING OF 1/2 WAVE PHASING SECTION. A 1/4 WAVE LINEAR TRANSFORMER OR A 1/2 WAVE STUB MAY BE USED TO MATCH THE FEEDLINE.



(B) 16 ELEMENTS
FEEDPOINT IMPEDANCE IS AROUND 300-400 OHMS, DEPENDING ON CONDUCTOR SIZE AND SPACING OF CENTER PHASING SECTION. A 1/4 WAVE LINEAR TRANSFORMER OR A 1/2 WAVE STUB MAY BE USED TO MATCH THE FEEDLINE. THE UPPER AND LOWER PHASING SECTIONS ARE TRANSPOSED. THEY MUST ALSO BE 1/2 WAVE LONG, BUT CONDUCTOR SIZE AND SPACING IS NOT CRITICAL.



(C) FEEDING 2 GROUPS WITH A MATCHING "HARNESS"

As an example of how the aperture formula (7) checks with the chart of Figure 10, let us take an array having 12 broadside and 8 colinear elements, or a total of 96 radiators, plus 96 reflectors. This antenna has an effective aperture six wavelengths high by four wavelengths wide, or 24 square wavelengths. (The effective aperture extends very slightly beyond the physical aperture on an array of this type). From formula (7), where: we have:

$$A_{eM} = 0.13\,G$$

$$24 = 0.13\,G$$

Thus: G = 185, or 22.6 db gain over a one half wave dipole. This agrees quite closely with the 22.4 db gain figure of the chart.

Extended Arrays

Extending the length of the elements from ½ wavelength to ⅝ wavelength, and increasing the broadside spacing to ¾ wavelength will increase the power gain in direct proportion to the greater aperture size. With smaller arrays this may be an advantage since an eight element array will show about as much gain as a conventional twelve element array. However, with larger systems, it is questionable that any advantage exists. For a given antenna size, the same amount of element material and phasing line must be used, and the supporting framework must be as large. Power gain will still be a measure of the effective aperture size, and the extended systems tend to produce stronger spurious side lobes.

The Yagi Array

The parasitic type end-fire antenna system, commonly called the *Yagi* because of Dr. Hidetsugu Yagi's pioneering work in this field, has a number of advantages for which it has gained prominent use in the VHF region. For a given power gain, the Yagi can be built lighter, more compact,

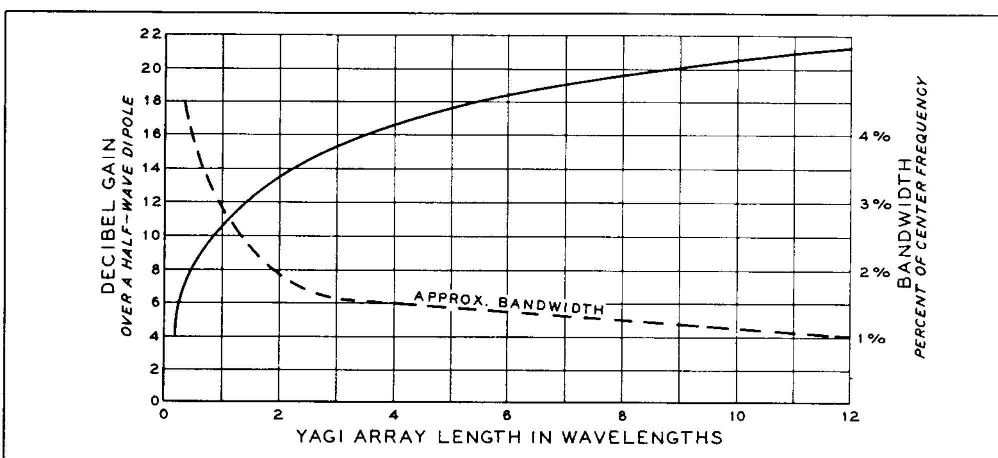


Fig. 11 Power gain and bandwidth of long Yagi antenna are expressed in terms of array length. Gain increases directly with length of the Yagi.

and with less wind resistance than any other antenna type. As a result, it can generally be erected to a greater height above ground than the more cumbersome arrays, thereby clearing obstructions more readily, and extending signal coverage. On the other hand, if a Yagi array of the same approximate size and weight as another antenna type is built, it will provide a higher order of power gain and directivity than that of the other antenna.

The price of this advantage is the relatively narrow operational bandwidth of the Yagi antenna when compared to arrays of other types. Yagi band width will depend upon the length of the array as well as on the number of arrays that are stacked and fed together. Figure 11 shows approximate bandwidth vs. Yagi length in wavelengths. The same graph also illustrates the power gain which can be achieved by a long Yagi. These curves represent an antenna which is adjusted for maximum forward gain. It is possible to design the Yagi antenna for greater bandwidth, but at a sacrifice in power gain.

It can be seen from the curve of Figure 11 that Yagi power gain increases directly with array length. The maximum practical length is entirely a problem of physically supporting the long series of director elements, although when the array exceeds a few wavelengths in length the element lengths, spacings, and Q's become more and more critical. The effectiveness of the director string depends upon a proper combination of the mutual coupling loops between adjacent directors and between the launching director and the driven element.

Practically all work on Yagi antennas with more than three or four elements has been on an experimental, cut-and-dry basis. A mathematical analysis of exactly what the various antenna dimensions should be has never been achieved successfully. This is due to the great complexity caused by the large number of variables in the problem, and also the practical fact that "the proof of the pudding is still in the eating thereof." With the increase in the use of long Yagi antennas, mathematicians will probably soon derive the formulas and proof of why these antennas work so well. In the meantime, we must rely upon the use of charts and graphs that have been derived from experimental data.

Element Spacing

There are a number of combinations for spacing between director elements which give good results. In general, optimum results are obtained when the launching director is placed quite close to the driven element, with gradually increased spacing out to director number five, after which a constant spacing of 0.39 wavelength is optimum. The first two or three Yagi directors are referred to as "launching directors" and serve to increase the coupling from the dipole to the wide spaced director string. The chart of Figure 12 gives parasitic element spacing which will produce good results with Yagi antennas up to 10 wavelengths long.

Reflector spacing is not critical, with values between 0.15 and 0.25 wavelength being satisfactory. The exact spacing will have a slight effect upon the reflector length, and on the radiation resistance of the driven element.

Reflector Length

Reflector length will be very close to a full half wavelength long. The

DIRECTOR NUMBER	SPACING FROM PRECEEDING ELEMENT, WAVELENGTHS	BOOM LENGTH FROM DIPOLE, WAVELENGTHS
1	.08	.08
2	.09	.17
3	.09	.26
4	.20	.46
5	.39	.85
6	.39	1.24
7	.39	1.63
8	.39	2.02
9	.39	2.41
10	.39	2.80
11	.39	3.19
12	.39	3.58
13	.39	3.97
14	.39	4.36
15	.39	4.75
16	.39	5.14
17	.39	5.53
18	.39	5.92
19	.39	6.31
20	.39	6.70
21	.39	7.09
22	.39	7.48
23	.39	7.87
24	.39	8.26
25	.39	8.65
26	.39	9.04
27	.39	9.43
28	.39	9.82
29	.39	10.21
30	.39	10.40

Fig. 12 Spacing of parasitic elements of the Yagi array is tabulated at right. Spacing gradually increases up to 0.39 wavelength at the fifth director, after which the value remains constant. Reflector spacing is not critical.

reflector length should be adjusted while measuring the back radiation of the array at a distance of ten or more wavelengths. This measurement should be made at the "design center" frequency. A F/B ratio as high as 30 to 40 db may be obtained at this frequency, although the figure may deteriorate to about 20 db at frequencies ½% either side of design center. Extra reflector elements mounted ¼ wavelength above and below the first will tend to hold the F/B ratio constant over a broader frequency range.

Director Length

It is commonly believed that the director elements of the Yagi antenna must gradually decrease in length as they progress away from the dipole. Some erroneous designs have even had the director lengths fluctuate erratically. Much of this confusion results from the fact that the designer started with a Yagi of three or four elements, and simply added more directors. Also, improper measuring techniques, combined with serious ground reflection can produce misleading answers. Actually a complete *system design* program of extended Yagi antennas reveals the interesting fact that the directors should all be the same length. In fact, when used with the element spacings listed in Figure 12, forward gain will be maximum under these conditions.

Figure 13 is a graph which may be employed to approximate proper director length in an array wherein all directors are to be the same length. Curves are given for elements of various diameters. As an example: Let us suppose we wish to design a Yagi array having a total number of 16 directors. The boom length will be 5.14 wavelengths (from Figure 12), plus the reflector spacing. Power gain will be about 60 (from Figure 11), or 17.8 db over a half wave dipole. Elements are .001 wavelength in diameter. From Figure 13 we find that the directors should be 0.438 wavelength long. This figure is a close approximation, and with each new design it is best

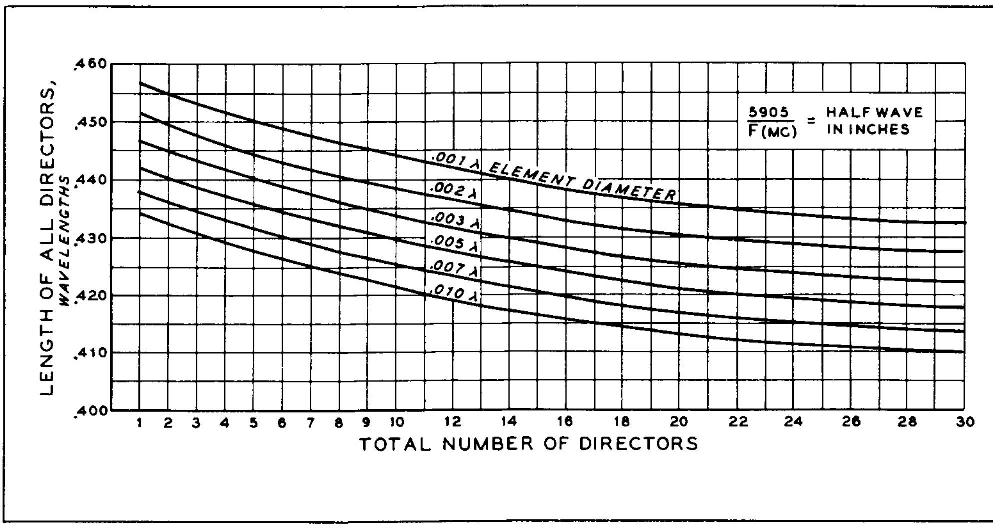


Fig. 13 Extended Yagi director dimensions for condition of optimum gain. All directors are of equal length and are spaced according to Figure 12. This graph is for insulated elements only. See text for grounded elements.

to check performance carefully, trimming element lengths from these findings.

The graph of Figure 14 will prove helpful in converting element diameters from wavelength measure to inches. Curves for various frequencies, including amateur bands, are included.

One of the best indications of proper element lengths in a long Yagi is the observation of side lobe strength. The frequency characteristic of an average array is shown in Figure 15, and shows how forward gain drops very sharply above the center design frequency, while side lobes come up rapidly. By setting up a method for measuring side lobes, it is possible to quickly determine the optimum operating frequency of the array. Side lobes should not be stronger than 10 to 12 db below the main lobe. As the

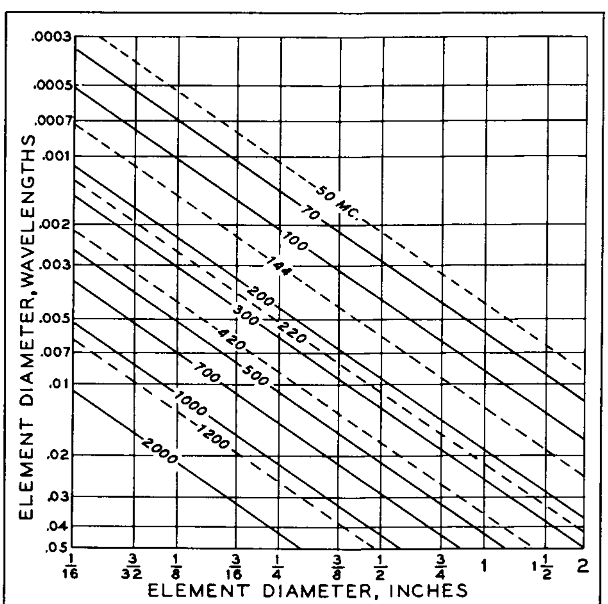


Fig. 14 Conversion chart for changing the element diameter from inches to wavelength measure.

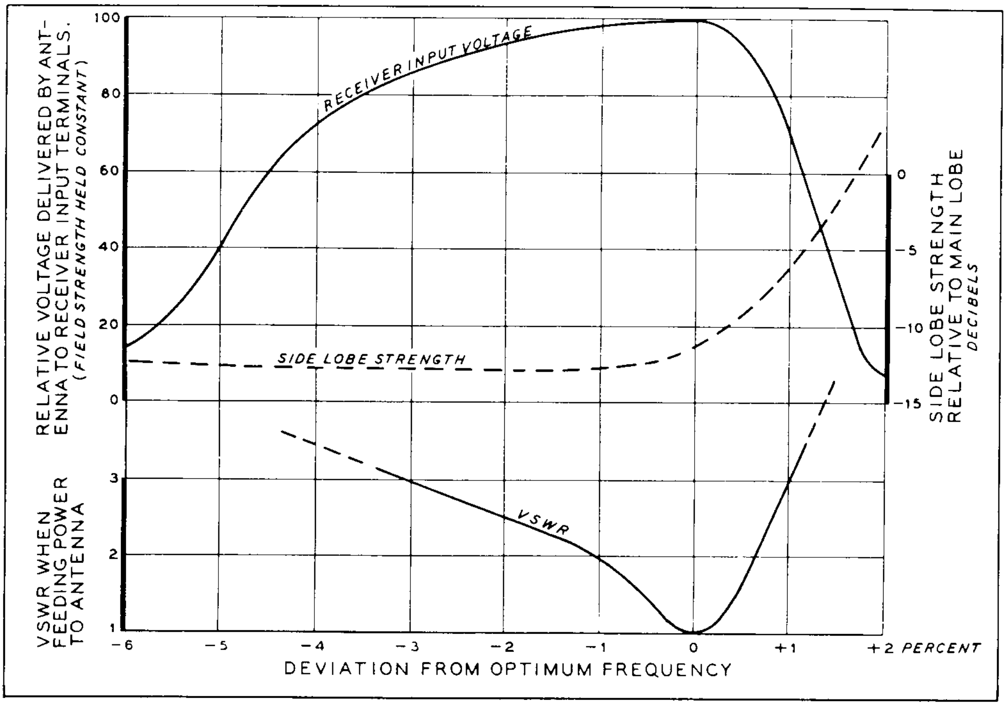


Fig. 15 Frequency characteristic of 4 wavelength Yagi. VSWR restricts the operational bandwidth. Gain is best on low frequency side of resonance.

operating frequency of the exciting signal is increased, a point is reached where the side lobes increase rapidly. Just below this frequency is the optimum operating frequency. If it is not the desired operating frequency, then element lengths must be changed accordingly.

The graph of Figure 13 can serve as a good starting point for the design of a long Yagi, but the one factor which is most difficult to predict is the effect of the supporting arrangement which holds each element. If the element center passes through a metal boom structure, an additive correction to the length of the element must be made. The exact correction figure depends upon the diameter of the element, and of the boom and is, unfortunately, hard to pin down since mounting hardware also enters into the picture. Generally, an amount equal to 2/3 of the boom diameter should be added to the length of each element. A final check of side lobe strength is highly recommended. If any doubt exists, it is better to cut the parasitic elements a trifle short as gain drops more slowly on the low frequency side of the optimum operating frequency, and side lobes decrease slightly, as shown in Figure 15.

Typical directivity patterns of a long Yagi antenna are shown in Figure 16. The generation of side lobes is a normal development of any multi-element antenna system, and although measures can be taken to reduce lobe size, it is practically impossible to eliminate them.

Staggered Director Lengths

If the extended Yagi is designed so that the directors gradually decrease in length as they progress from the dipole bandwidth will be increased, and both side lobes and forward gain will be reduced. As the degree of

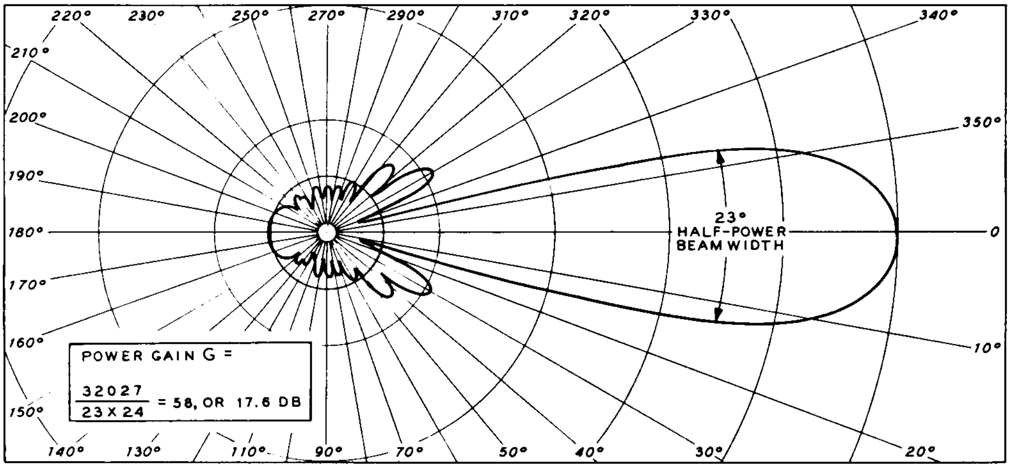


Fig. 16 Typical directional pattern of 5 wavelengths Yagi. Polarization is horizontal, and half-power beam in vertical plane is 24 degrees.

director length stagger is increased, bandwidth will become greater and forward gain less, until finally the advantages of the parasitic array disappear. Figure 17 illustrates design parameters for a wide-band staggered director array. Element spacings are the same as those listed in Figure 12.

One advantage gained from staggered director length is that the array can be lengthened or shortened by adding or taking away elements without the need for returning or correcting the original group of parasitic elements (or remaining group, as the case may be). When all directors are the same length, they must all be shortened *en masse* as the array is lengthened, and vice-versa when the array is shortened.

Element Thickness

The subject of element thickness, or diameter, in Yagi arrays is not clearly understood, and some disagreement exists as to the part played by the element thickness. Generally speaking, the larger diameter element will exhibit a broader bandwidth characteristic than the element of smaller diameter. This can work to advantage when the element is employed as a radiator in a wide-band antenna system. This fact may lead to the assumption that the operating bandwidth of a Yagi antenna can be increased by employing thicker parasitic elements. However, the coefficient of coupling between the directors of a long, extended Yagi array is extremely critical, and it becomes necessary to move the thicker parasitic elements closer together in order to achieve maximum forward gain.

Very little unclassified information is available on the subject, and so in this case it is necessary for the authors to draw conclusions based upon their own experimental observations. In all cases it has been found impossible to secure as high a forward gain figure with thick elements as with thin ones. All indications are that antenna gain will be highest when employing the thinnest possible parasitic elements, until the point is reached where the surface conductivity becomes poor enough to materially affect the Q of the element.

For this reason it seems probable that the Yagi array cannot be scaled directly to a higher frequency and produce the same results that may be measured at a lower frequency, since the element thickness will have to

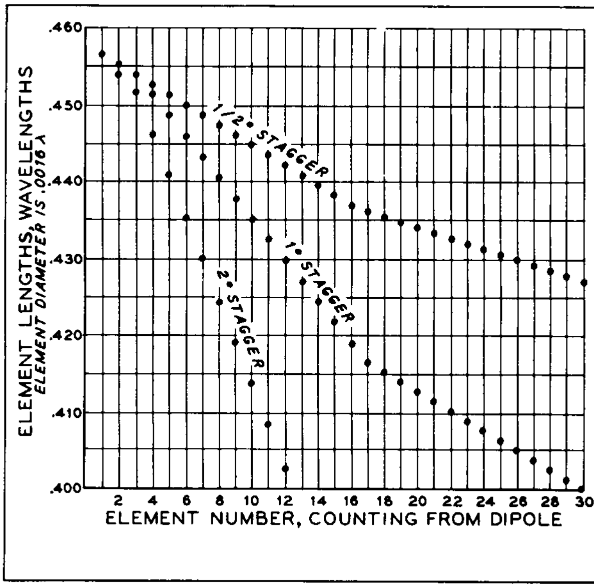


Fig. 17 Director dimensions for extended Yagis having staggered director lengths. The elements are spaced according to the chart of Figure 12. Three degrees of stagger are shown. Forward gain drops with increased stagger, while bandwidth becomes greater.

be greater at the higher frequency for purely mechanical reasons.

Due to r-f "skin effect", conductivity at the surface of the small diameter elements is a critical factor. It is important that the best practical material be employed for parasitic elements. Silver is well known as the best conductor, however if the element is to be silver plated the coating must be properly applied as a smooth, homogeneous plating. A poor silver plating job may offer more resistance to r-f currents than plain copper or aluminum. Since silver oxide is also a good conductor, the silver plating may be allowed to oxidize without affecting antenna performance. Other metals are not as fortunate, however. If copper or aluminum are allowed to oxidize their surface resistance will increase considerably. They therefore should be protected by a thin coating of polystyrene or a similar substance having high insulating properties and low dielectric constant.

One of the most practical plating materials is gold, since it will not oxidize and it provides very good conductivity. It can be applied in extremely thin layers, and although the cost is somewhat higher than that of silver plating, there is less chance of an inferior plating job.

Radiation Resistance of the Extended Yagi

The radiation resistance of the driven element in an extended Yagi antenna turns out to be somewhat higher than would normally be expected. The value of radiation resistance rises and falls as directors are added, but the median value of resistance is about 20 ohms. Radiation resistance values smaller than this are sometimes encountered with shorter arrays having only one or two directors. This lower value in the short arrays may be due to serious overcoupling between the parasitics and the driven element.

Stacking Yagi Antennas

Any number of antennas may be stacked in vertical or horizontal rows and coupled together to provide additional gain and directivity. It is important, however, that such antennas be spaced correctly between themselves in order to realize all possible gain. This is where a knowledge of the

effective aperture size of the individual antennas becomes useful. By spacing the antennas so their effective apertures just "touch", power gain will increase directly as the number of antennas used. In other words, stacking four antennas will increase the gain four times (6 db) over that of one antenna, etc. If the antennas are spaced at insufficient distance from each other, the effective apertures will overlap and the power gain of the structure will suffer.

The directional pattern of the array will change according to the direction of stacking. If the array will change according to the direction of stacking. If the array is made four antennas wide, the beamwidth in the horizontal plane will be one-fourth of the beamwidth of one antenna. If the array is made two antennas high, the vertical beam width will be half that of one antenna alone. Applied as a formula, we have:

$$\theta_{EA} = \frac{\theta_{ES}}{N}$$

THE SAME FORMULA APPLIES IN THE H-PLANE.

WHERE: θ_{EA} IS BEAM WIDTH OF ARRAY IN E-PLANE

θ_{ES} IS BEAM WIDTH OF SINGLE ANTENNA IN E-PLANE

N = NUMBER OF ANTENNAS STACKED IN E-PLANE

As a simple rule of thumb in stacking extended Yagi antennas, it is recommended that they be spaced a distance equal to ¾ of the length of the Yagi. This figure will be quite close to the aperture size of the individual antenna.

Naturally, proper phasing and impedance matching is essential in coupling a group of antennas to one transmission line. Since extended Yagis must be placed quite far apart in terms of wavelength, it is usually the best practice that the feedline running from each antenna be made of a characteristic impedance which matches the individual radiator, or the matching device located at the radiator. In other words, if each antenna is designed to match a 52 ohm line, it is best to use 52 ohm line from each antenna to the common junction point of the feed system. At the junction the necessary transformers or baluns can be applied to match the feedlines to the main transmission line. Refer to actual construction data given later in this Handbook for specific examples.

Matching Systems for Yagi Antennas

The type of matching system to be used with Yagi antennas depends upon the type of feeder system to be employed. If a balanced, open wire transmission line is to be used, the folded dipole is recommended (Figure 18 A). If a shielded, coaxial transmission line is employed, the unbalanced Gamma match system (Figure 18 B) should be used. Complete information covering the design and adjustment of these systems is given in the *Beam Antenna Handbook*, published by Radio Publications, Danbury Road, Wilton, Conn.

The purpose of these matching systems is to transform the low radiation resistance of the Yagi antenna to a value equal to the characteristic impedance of the transmission line. Under conditions of proper match, the standing wave ratio (SWR) on the transmission line will be close to 1/1 over a narrow frequency range either side of the resonance point of the driven element of the array.

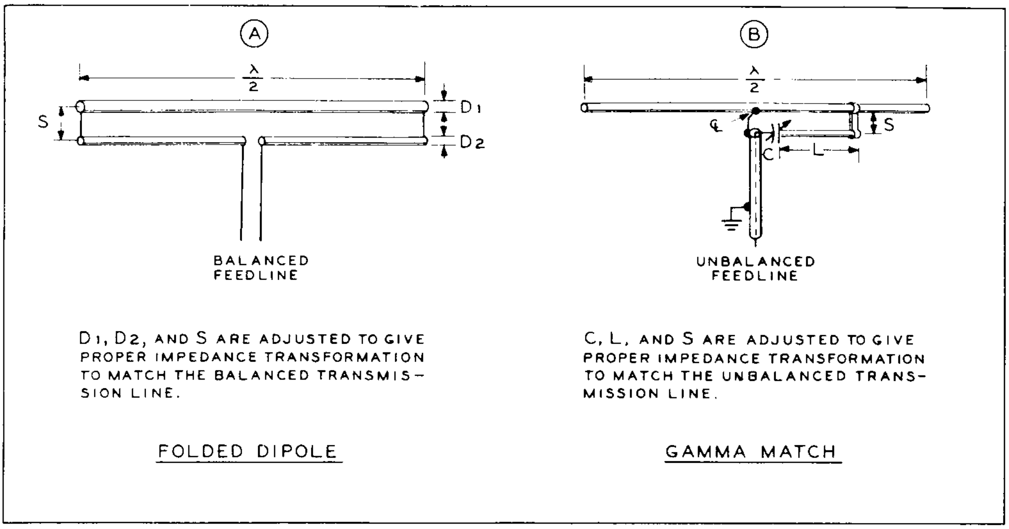


Fig. 18 A folded dipole may be used for the driven element of a Yagi array if a balanced feedline is employed. The ratio of impedance transformation is determined by the diameter ratios of the sections of the dipole (D), and the spacing between the dipole elements (S). An unbalanced matching system such as the Gamma match should be employed when the antenna is fed with coaxial transmission line. Transformation ratio is set by length of gamma rod (L) and element spacing (S). System is resonated by series capacitor (C).

Other Antenna Types

Within the limited space of this text it is impossible to discuss all possible antenna types. Thus far we have covered the two types most commonly used by the radio amateur, namely, the colinear-broadside array, and the Yagi array. The following is a brief summary of some other antenna types occasionally used in the VHF region.

Long Wires, V-beams, Rhombics

These antennas are characterized by high gain and generally broad frequency bandwidth. Usually they are too unwieldly for rotation, although in UHF TV work some compact arrangements have proven to be successful. These antennas are used for point to point work where orientation need not be changed, and are also used for VHF scatter work. All of these antennas are critical as to the ground conductivity directly below the array, and they should be used over flat ground having good conductivity characteristics.

Corner Reflectors

These antennas feature a high front to back ratio and good bandwidth characteristics, along with good forward gain. A 90-degree corner will provide 10 db gain, a 60-degree corner nearly 12 db, and a 45-degree about about 14 db. The power gain is produced almost entirely by compression in the H-plane, and for this reason the horizontally polarized corner reflector does not give outstanding results on long paths beyond the horizon where a wide vertical acceptance angle can be very beneficial. In the UHF region the corner provides a simple, high gain antenna for point to point service.

CHAPTER VII

VHF Antenna Construction

Regardless of the type of VHF antenna to be built, the strongest, lightest and easiest type of construction is the all-metal assembly. This configuration employs a metal boom, with all parasitic elements fastened directly to the boom. The driven elements may or may not be insulated from the boom, depending upon the feed system used with the antenna.

A complete discussion of antenna construction is contained in the *Beam Antenna Handbook*, distributed by the publishers of this Handbook. The reader is referred to this text for general information covering antenna construction. This information applies equally well to VHF antennas. However, certain constructional points peculiar to VHF antennas will be discussed in this chapter. In particular, the long Yagi type antenna requires special assembly techniques to achieve best operational results.

Mechanical Assemblies

It is fortunate that much construction material for VHF arrays is available in the form of TV-type hardware. Mast sections, elements, and connecting joints may be made up of standard TV antenna parts obtainable at the larger radio and television parts distributors. Shown in Figure 1 are three methods for joining mast and support sections. These assemblies may be used to attach the antenna boom to the supporting structure.

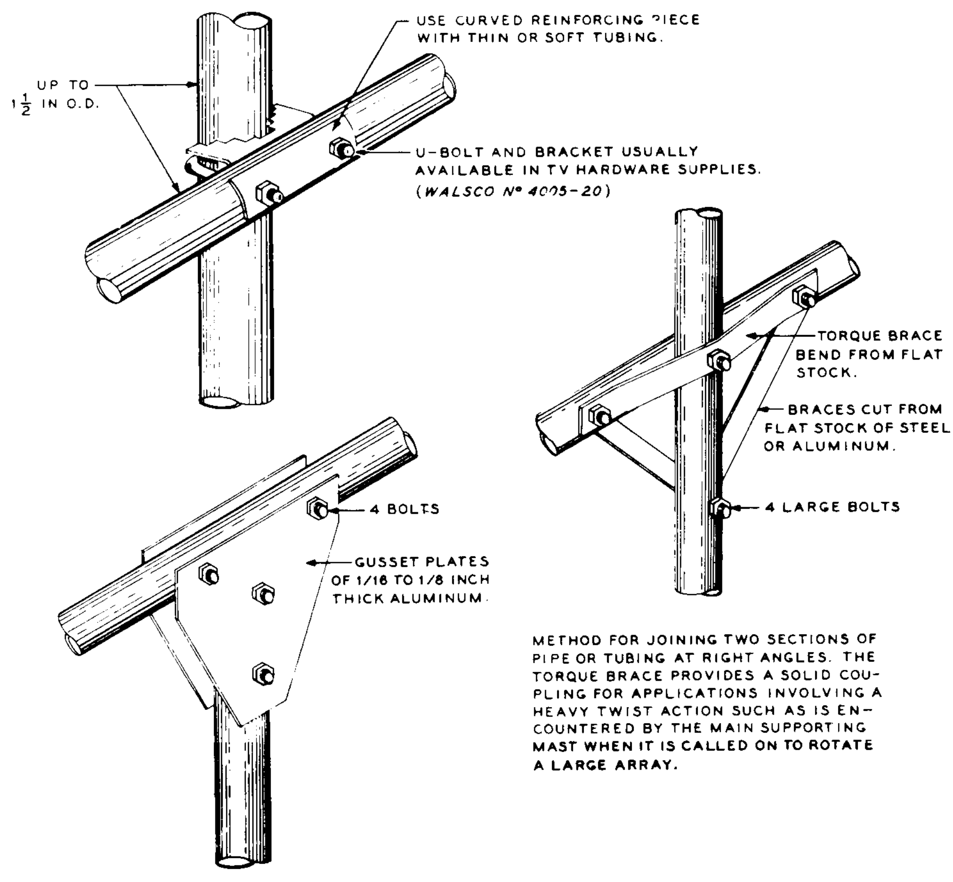
The radiating and parasitic elements of VHF arrays are generally of smaller diameter than the boom and present a problem in support and mounting. Figure 2 shows four different methods of mounting elements to a metal boom. Methods B and D are especially effective for mounting thin parasitic elements to a metal boom. By employing the assemblies of Figures 1 and 2, VHF arrays of any size may be built, with the assurance that the array will have sufficient mechanical strength to withstand strong winds, and inclement weather which would mark the end of flimsy antenna structures.

VHF Linear Matching Transformers

Because of the relatively small size of the VHF elements, the folded dipole is a popular form of radiator. The parameters of the dipole are usually

FIGURE 1

METHODS FOR JOINING MAST AND SUPPORT SECTIONS



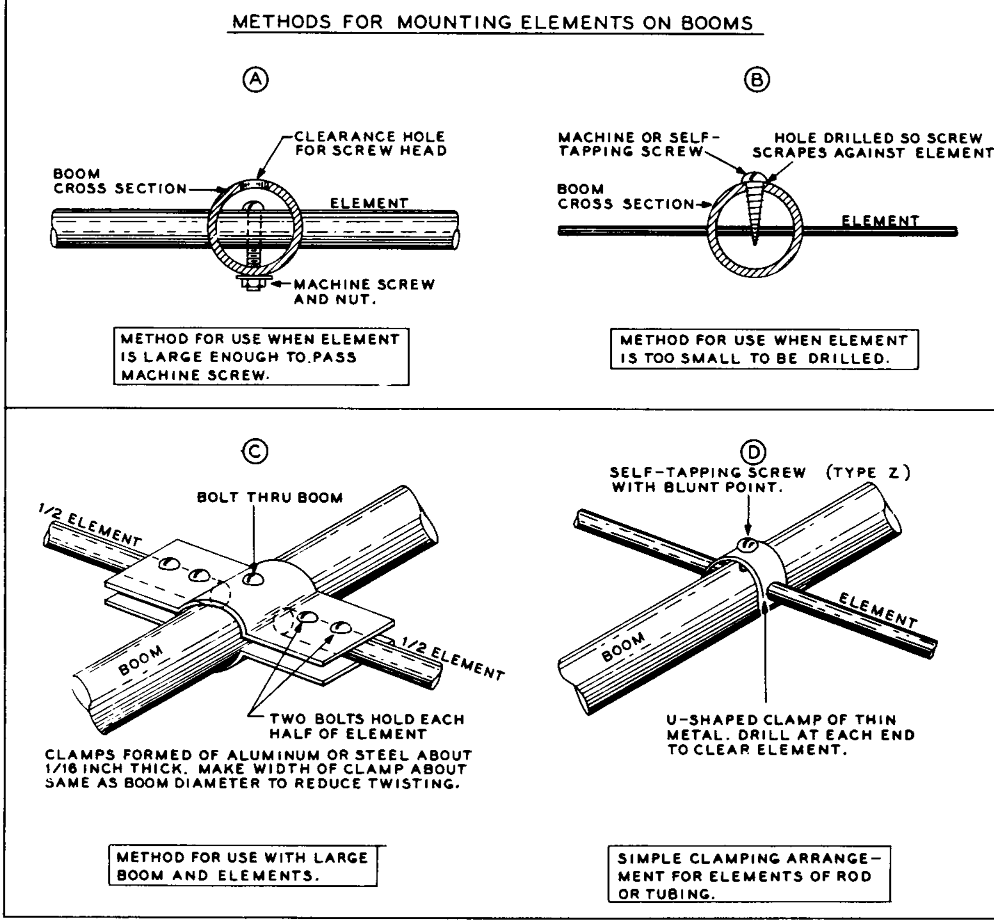
METHOD FOR JOINING TWO SECTIONS OF PIPE OR TUBING AT RIGHT ANGLES. THE TORQUE BRACE PROVIDES A SOLID COUPLING FOR APPLICATIONS INVOLVING A HEAVY TWIST ACTION SUCH AS IS ENCOUNTERED BY THE MAIN SUPPORTING MAST WHEN IT IS CALLED ON TO ROTATE A LARGE ARRAY.

ARRANGEMENT WHICH MAKES AN EFFECTIVE T-JOINT WHEN TUBING IS OF SAME DIAMETER.

chosen so that it presents a 300 ohm or 450 ohm termination for the transmission line. When an array of arrays is used, or when it is desired to employ a low impedance transmission line from the antenna to the transmitter, it is necessary to transform impedances with the use of a linear matching transformer, as shown in Figure 3. Drawing A illustrates a balanced transformer which will transform one impedance to another value, as determined by the configuration of the transformer. A transformer for unbalanced, coaxial systems is shown in drawing B. The outer shell of the coaxial transformer may be grounded, but it is necessary to keep the balanced transformer well in the clear, and out of strong r-f fields to insure proper operation.

A second type of impedance matching device is the *Balun,* which is a matching device used when connecting a balanced termination to an unbalanced one. The balun may provide this action at 1:1 transformation ratio,

FIGURE 2



as shown in Figure 4A, or it may be constructed to provide a 4:1 impedance transformation, as shown in Figure 4B. The simple baluns shown are made of sections of coaxial transmission line, and are satisfactory where an inexpensive device is desired for use at 50 mc or 144 mc. For proper operation at the higher frequencies, however, a more precise type of instrument is required, since there is a small degree of inherent unbalance in the devices of Figure 4, no matter how well built. In addition, the r-f loss of these solid dielectric designs increases with frequency. Highest efficiency and best balance to ground may be obtained with an air-dielectric balun, such as the one illustrated in Figure 5. This type of construction will perform well into the upper limits of the UHF region, and its use is recommended where the utmost efficiency and circuit balance is desired. For use above 220 mc, the SO-239 connector should be replaced with the UHF type UG-58/U constant impedance connector. A minimum number of spacers of the highest quality should be used to preserve the rigidity of the assembly.

Matching Harnesses for Feeding Groups of Arrays

One of the most perplexing problems is that of applying power to the

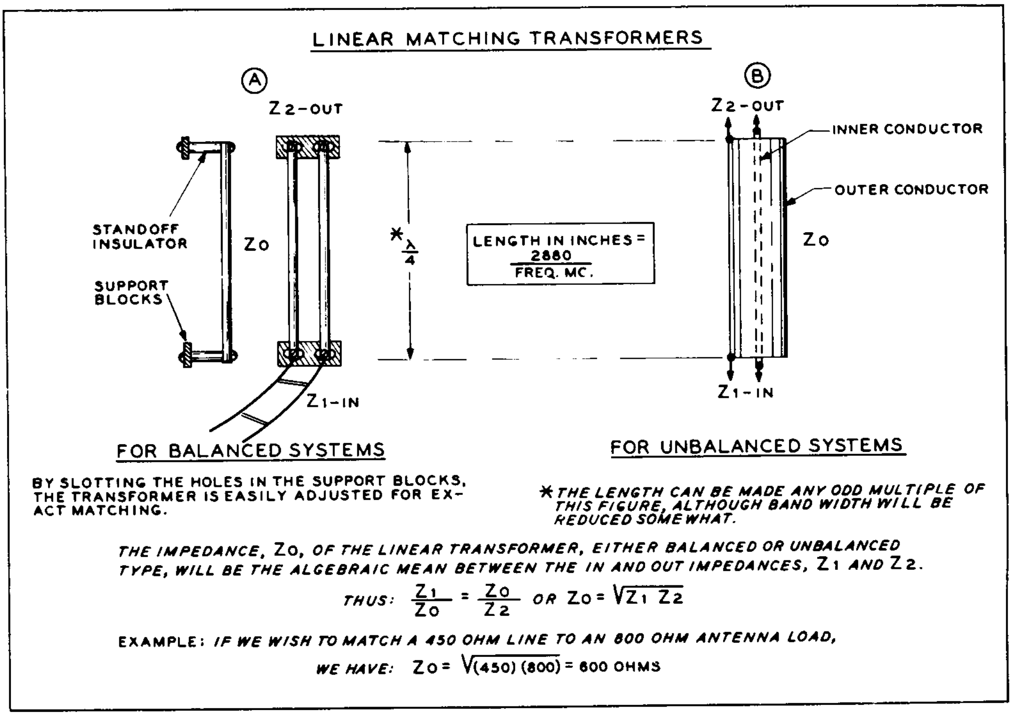


Fig. 3 Linear matching transformers are used to change value of antenna impedance to match the characteristic impedance of the transmission line.

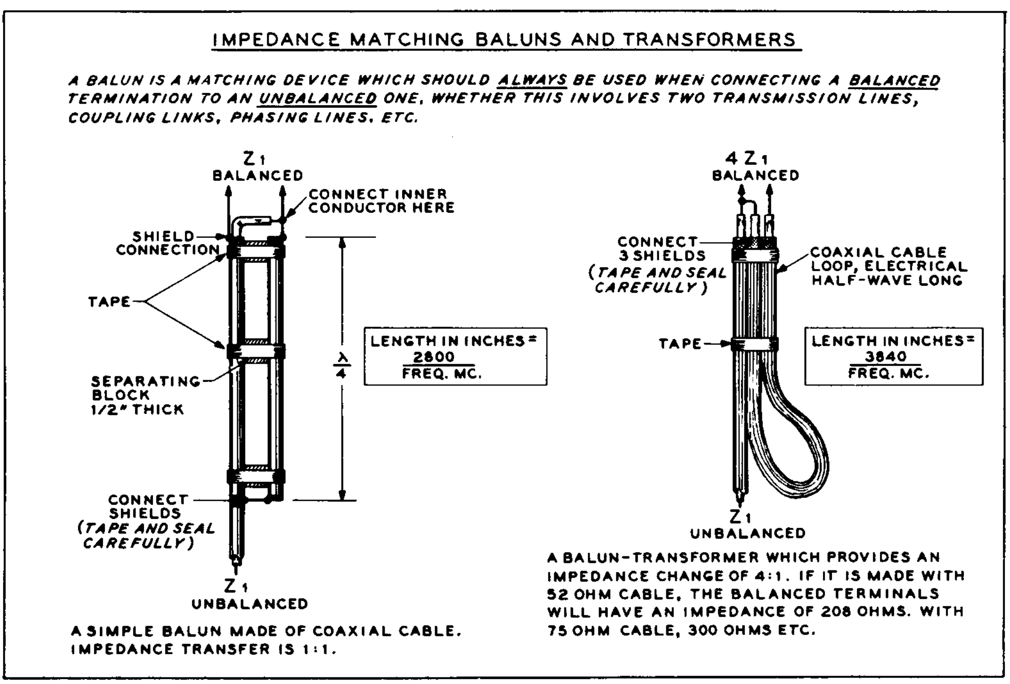
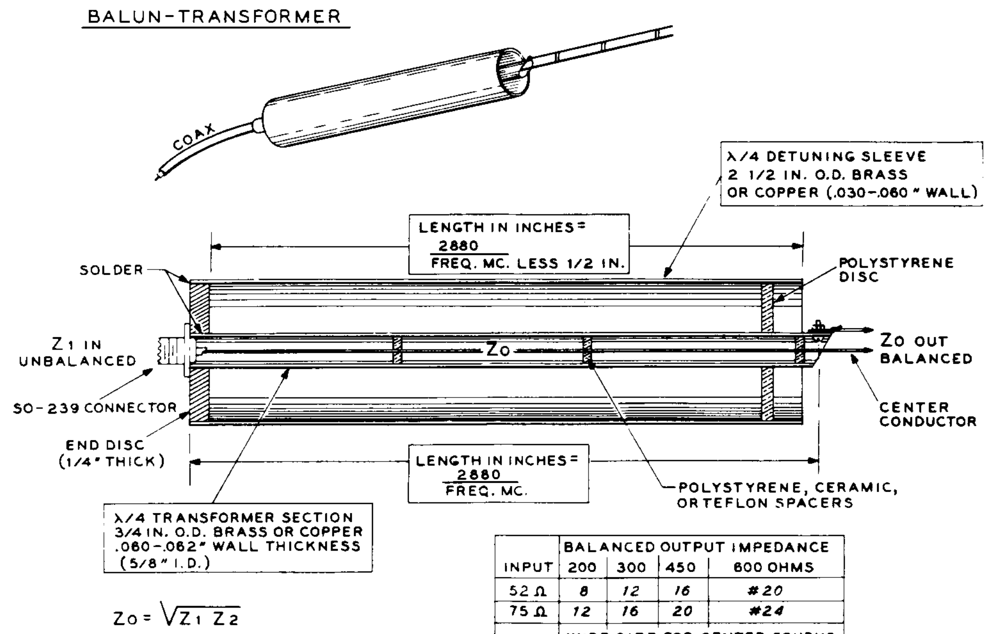


Fig. 4 The Balun is a matching device used to connect a balanced termination to an unbalanced one. Baluns must be tuned to the frequency of use.



	BALANCED OUTPUT IMPEDANCE			
INPUT	200	300	450	600 OHMS
52 Ω	8	12	16	#20
75 Ω	12	16	20	#24
	WIRE SIZE FOR CENTER CONDUC.			

ILLUSTRATED ABOVE IS A LINEAR BALUN-TRANSFORMER WHICH IS NOT DIFFICULT TO CONSTRUCT, AND PROVIDES HIGHER EFFICIENCY THAN SOME OF THE MORE SIMPLE TYPES. THE PARTS MAY BE SILVER PLATED FOR MAXIMUM CONDUCTIVITY. IF THE BALUN IS TO BE USED OUTDOORS, MOUNT IT VERTICALLY WITH THE COAX CABLE COMING OUT THE TOP IN ORDER TO KEEP THE RAIN OUT. TAPE AND SEAL THE MALE CONNECTOR TO THE BALUN AFTER INSTALLATION.

Fig. 5 Air dielectric balun provides greatest degree of balance at VHF.

driven elements of a group of arrays in a manner assuring equal division of power between the elements. The secret of success is the use of a completely symmetrical feed system, such as is shown in Figure 6. Three examples are given, illustrating how a manifold feed system may be developed for feeding any number of driven elements. There are innumerable arrangements which may be used, and the one described is only one of them. However, it is a fairly simple system and will function well without critical adjustments. The following rules will insure good impedance matching and phasing:

1—Use balanced phasing lines in the harness, such as twin-lead, open wire line, or paralleled coaxial line. A balun may be employed at the master feedpoint (F) if coaxial cable is to be used for the main feedline (L4).

2—Make the phasing lines (L1, L2 and L3) all the same impedance and of the same material.

3—Lines L1, L2, and L3 may be any convenient length, although it is desireable that they be no longer than necessary. All L1 lines must be exact equal length, all L2 lines the same, and all L3 lines equal length. No relationship need be held between the lengths of the L1's as compared to the L2's and the L3's, however.

4—The antennas must be identical and well matched to the impedance of the phasing line, L1. Conduct tests and adjustments with one antenna alone until this condition is met. In other words, if 450 ohm

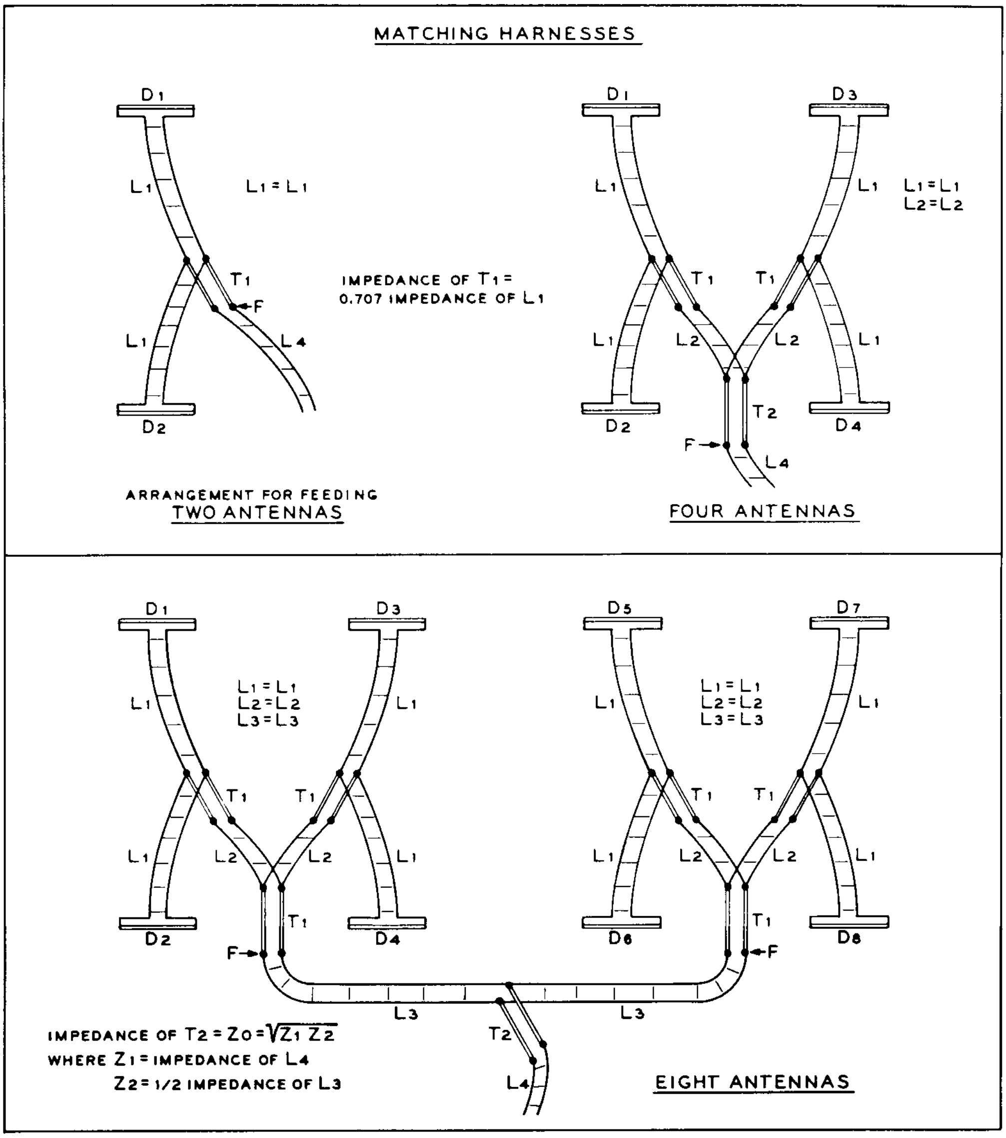


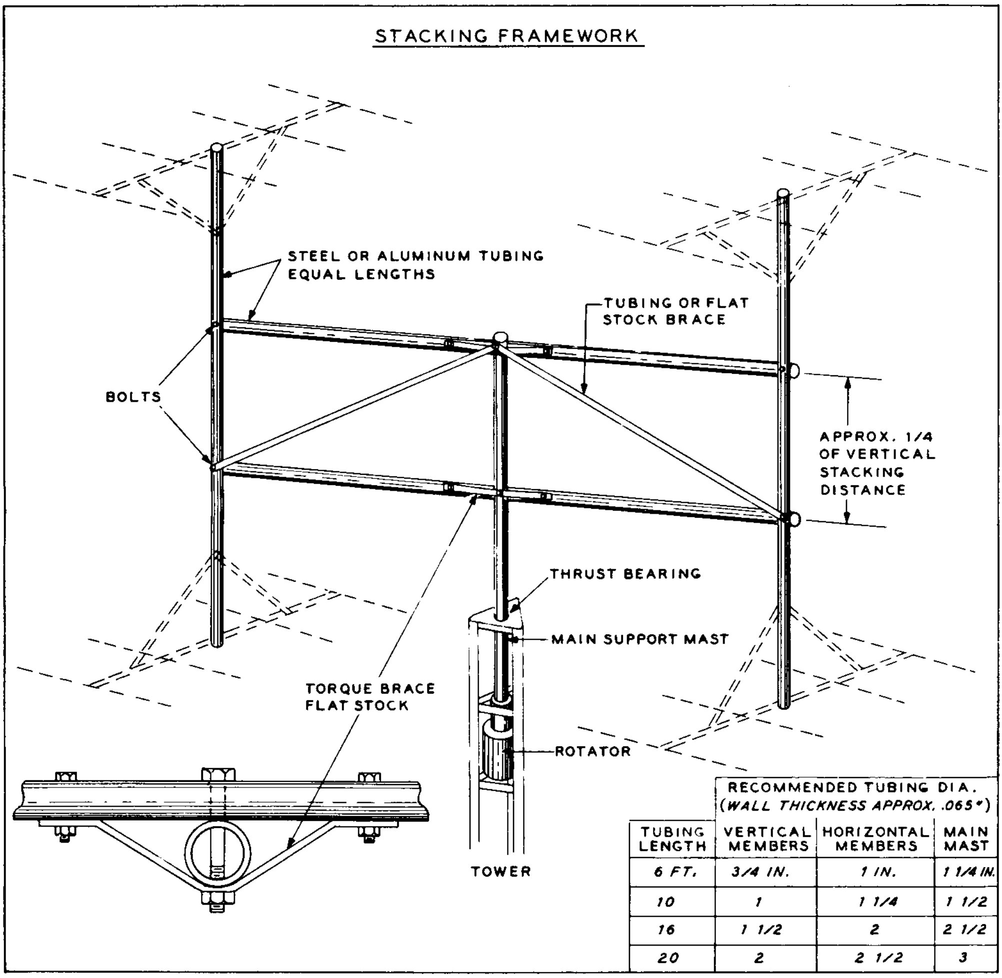
Fig. 6 Symmetrical matching harnesses are used to drive in-phase elements.

phasing lines are used in the manifold harness, each individual antenna must be adjusted for a 450 ohm termination.

5—The T1 sections are ¼ wave linear transformers, and should be constructed so as to have a characteristic impedance equal to 0.707 of the impedance of the phasing lines used. For example, if 450 ohm phasing line is employed, the T1 sections should be 318 ohms. Refer to the *Beam Antenna Handbook* for additional information about linear transformers.

6—Harness polarity must be observed to maintain proper phasing. Simply make certain that the left hand terminal of all the driven elements (D) are all tied together by the harness, and likewise the right hand terminals.

7—The T2 section is a ¼ wave linear transformer whose characteristic impedance depends upon the impedance of the main feedline, L4. If



TUBING LENGTH	RECOMMENDED TUBING DIA. (*WALL THICKNESS APPROX. .065"*)		
	VERTICAL MEMBERS	HORIZONTAL MEMBERS	MAIN MAST
6 FT.	*3/4 IN.*	*1 IN.*	*1 1/4 IN.*
10	*1*	*1 1/4*	*1 1/2*
16	*1 1/2*	*2*	*2 1/2*
20	*2*	*2 1/2*	*3*

Fig. 7 Simple metal framework construction to support four long Yagi antennas.

L4 is of the same impedance as the phasing lines, then T2 will be identical to transformer T1. Otherwise, T2 should be calculated from the formula given in Figure 6.

8—Provide frequent support for the lines, and keep them clear of the antennas and of one another as much as possible. At junction points the lines should separate in different directions, rather than running parallel to one another.

The Stacking Framework

It is of utmost importance that a suitably rugged structure be provided to support the antenna array and manifold harness. Figure 7 illustrates the construction of a frame that will support four long Yagi antennas, or other equal types. Refer to the graph of Figure 8 for the stacking distance required between the Yagis. This will determine the dimensions, tubing diameters, etc. of the framework.

Very often in the design and construction of a large framework of wood or metal too little consideration is given to the lateral forces which a heavy wind will exert on the structure. This thrust is a function of the frontal area

facing the wind, and its distance away from the supporting mast. The downward pull of gravity is a relatively small force compared to that of a strong wind, so the lateral bracing of the framework deserves at least as much attention as the vertical bracing. A good rule of thumb to follow is to design the outer members of the assembly with adequate strength, and then increase the size of successive members as the design progresses inwards toward the main supporting mast. Needless to say, the main mast should be the largest and strongest member of the configuration.

Practical Antenna Assemblies

The remainder of this chapter is devoted to assembly information for practical, high gain arrays for the VHF bands. These antennas may be built from the graphs and drawings with the assurance that they will perform in an excellent manner without the need of further adjustment, provided they are erected well in the clear, with no large metal bodies nearby to exert detuning effects. The antennas are designed to be used with a balanced feed system. If coaxial feed is desired, the balun assemblies shown in Figures 4 and 5 should be used. The balun transformer should be mounted to the supporting mast a foot or so under the antenna array, and a short length of balanced transmission line of the proper characteristic impedance run from the balun to the driven element of the array.

50 Mc Yagi—9 Elements on a 30 Foot Boom—12.6 DB Gain

Shown in Figure 9 is the "ultimate" antenna for 50 mc. Although a 30-foot Yagi may appear to be a bit on the large side, it is really no more to handle than a 2 element 20 meter beam. The 12.6 db of actual power gain on 6 meters will produce outstanding results, especially on iono-

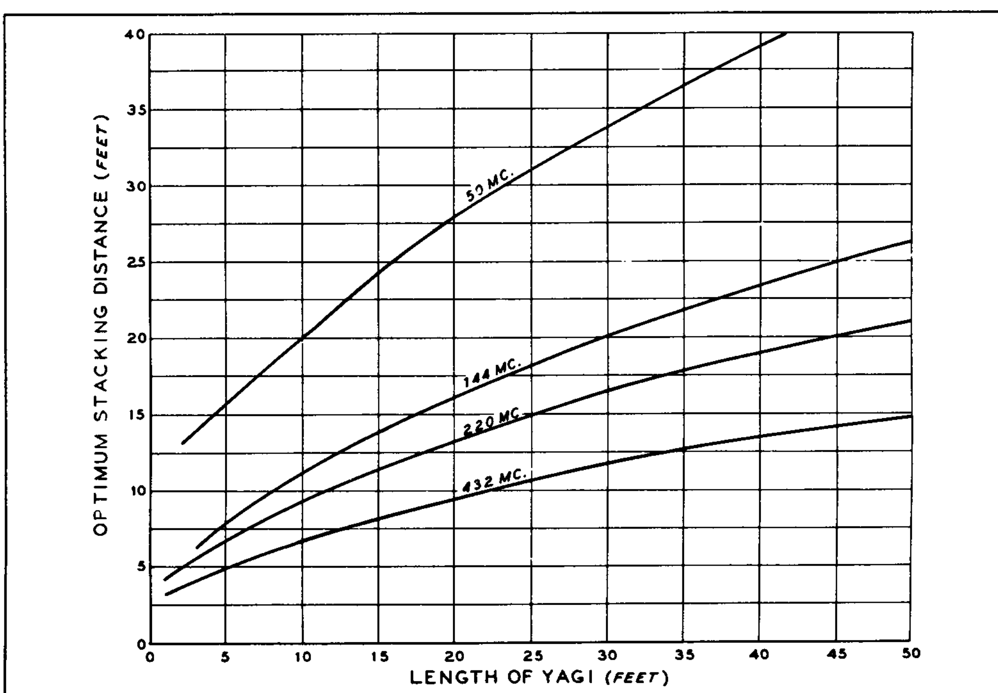
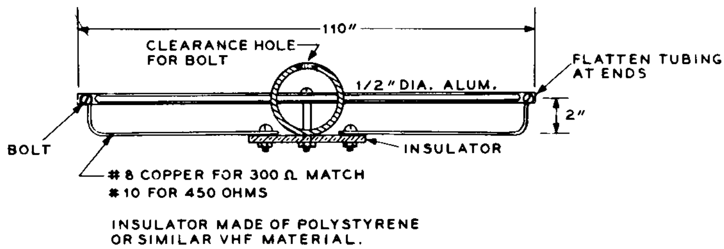


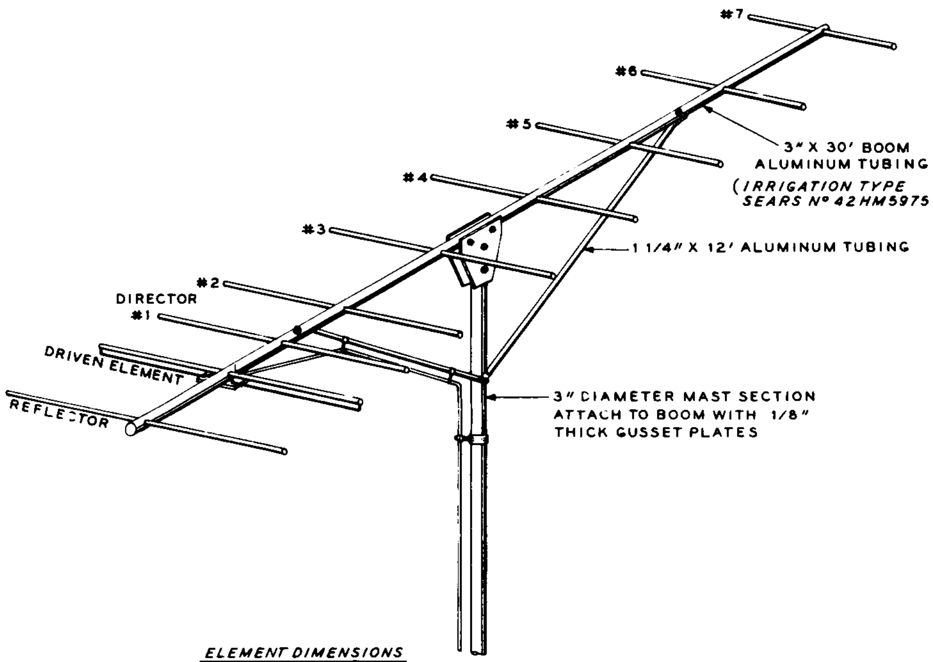
Fig. 8 Optimum stacking distance of long Yagi increases with length of boom.

50 MC YAGI---9 ELEMENTS ON A 30 FOOT BOOM---12.6 DB GAIN

FIGURE 9

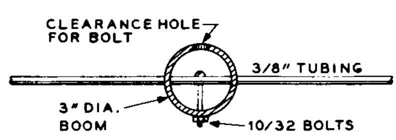
DRIVEN ELEMENT CONSTRUCTION





ELEMENT DIMENSIONS

	LENGTH 50 MC.	LENGTH 51 MC.	SPACING FROM DIPOLE
REFLECTOR	117″	114 3/4″	48″
DIRECTOR 1	105	102 3/4	23
2	104	101 3/4	71
3	103 1/4	101	119
4	102 1/2	100 1/4	167
5	101 3/4	99 1/2	215
6	101	98 3/4	263
7	100 1/4	98	311



METHOD FOR ATTACHING PARASITIC ELEMENTS TO BOOM

spheric scatter circuits. For those who would go so far as to stack two or more of these antennas, the proper stacking distance is 30 to 34 feet.

Staggered director elements are used in this design. The maximum gain figure drops about one db, but the array is now capable of operation from 50 to 51 mc when it is peaked at 50 mc. By cutting the elements for 51 mc as shown in the chart, performance will be good from 50 to 52 mc, with a slight drop in gain at 50 mc. This choice is left to the constructor. The staggered elements also allow the beam to be shortened without dis-

Fig. 10 Design data for staggered director Yagi antenna applies also to shorter Yagi, as shown in chart at right. Gain drops slowly as the boom length decreases.

NUMBER OF ELEMENTS	BOOM LENGTH	GAIN DB
9	30	12.6
8	26	12.0
7	22	11.2
6	18	10.4
5	14	9.4
4	10	8.3

turbing the remaining elements. The above design information, therefore, will apply to smaller antennas, as shown in the chart of Figure 10.

All parasitic elements are made of 3/8″ diameter 24ST or 61ST aluminum alloy tubing, having a wall thickness of .030″ to .060″. A thin coating of lacquer on the elements is advisable in order to delay the effects of oxidation.

144 Mc Yagi—10 Elements on a 12 Foot Boom—13.4 DB Gain

The medium length Yagi shown in Figure 11 provides good forward gain, yet is extremely light in weight with low wind resistance. Front-to-back ratio is about 30 db when the antenna is erected well clear of surrounding objects. The director lengths are staggered in order to broaden the frequency range and allow operation from 144 mc to about 147 mc without serious transmission line mismatch. Proper stacking distance of this array is 10 to 12 feet. If coaxial cable is to be used for the transmission line, a balun transformer should be mounted on the supporting mast.

All parasitic elements are made of 1/8″ diameter rod, preferably silver plated steel, or brass. 24ST aluminum alloy is satisfactory, but it should be treated with a thin coating of protective lacquer after assembly to retard oxidation.

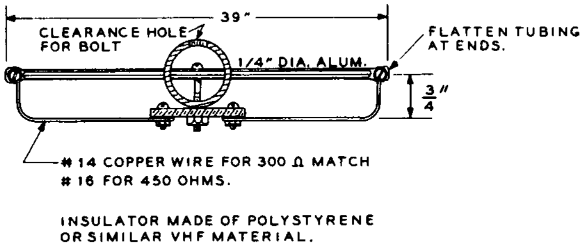
If it should be desired to mount this antenna for vertical polarization, the upper four feet of the supporting mast should be made of wood. The braces of the boom should also be wood in order to keep the electric plane of the antenna free of metal. The most convenient arrangement for vertical polarization may be to stack two antennas side by side with a horizontal supporting section which is held in the center by the main mast. These comments also apply to vertical mounting of the other Yagi antennas described in these pages.

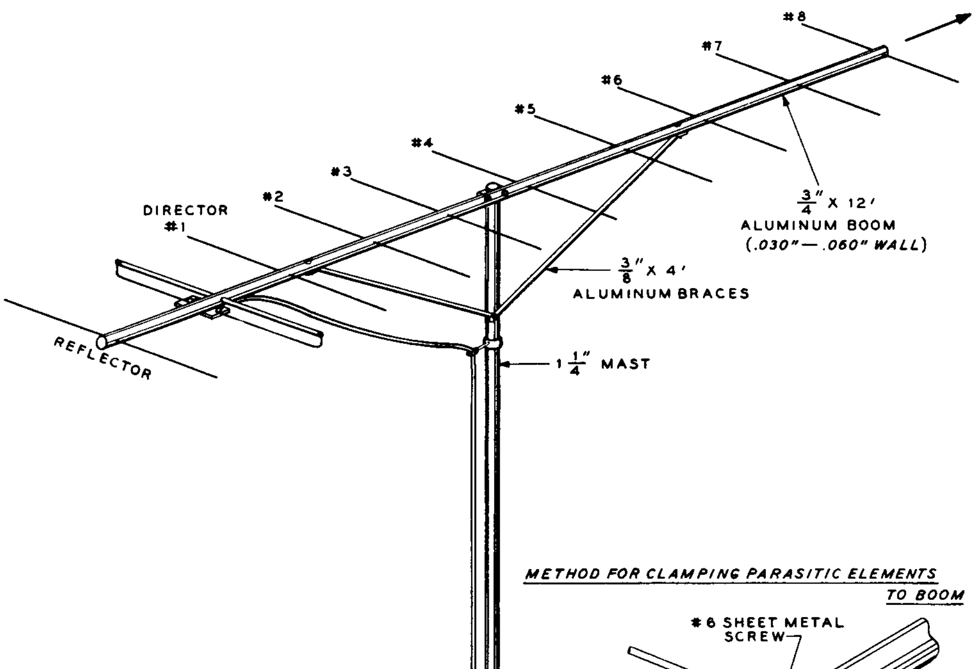
144 Mc Yagi—13 Elements on a 24 Foot Boom—16.1 DB Gain

This "ultimate" Yagi employs directors of uniform length, providing a forward gain figure of 16.1 to 16.4 db over a half wave dipole (Figure 12). Bandwidth extends from 1 mc above, to 1.5 mc below the design center frequency. Within these limits the VSWR will be less than 2/1. If the operator wishes to have maximum gain at 144 mc, the antenna will not operate well above 145 mc. On the other hand, if design center is placed at 145 mc, forward gain will be down about 1 db from maximum at 144 mc. The particular application of the antenna must determine the choice. The length of this array, and the total number of elements cannot be changed without reverting to staggered director lengths, as explained earlier.

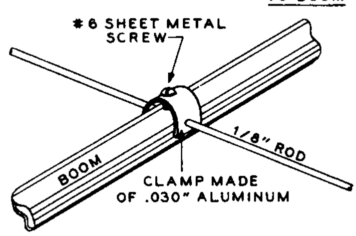
144 MC. YAGI --- 10 ELEMENTS ON A 12 FOOT BOOM --- 13.4 DB GAIN

DRIVEN ELEMENT CONSTRUCTION





METHOD FOR CLAMPING PARASITIC ELEMENTS TO BOOM

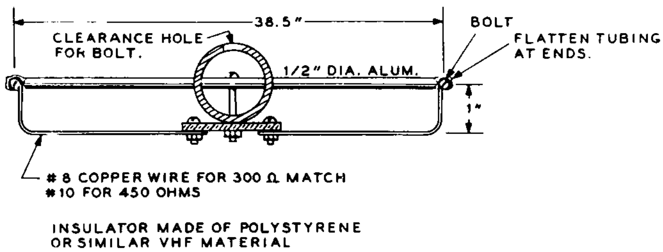


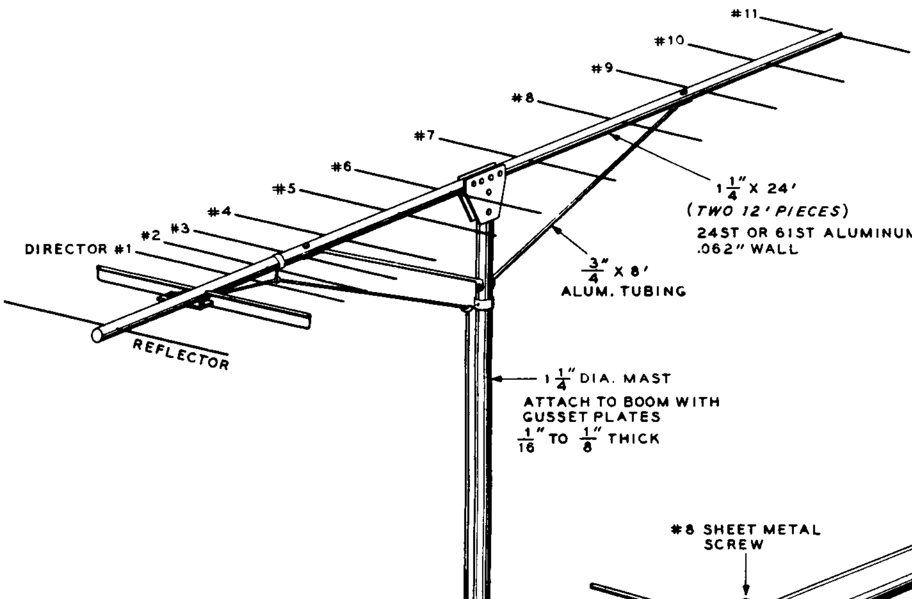
ELEMENT DIMENSIONS

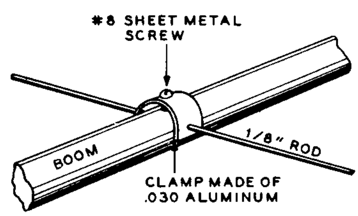
	LENGTH	SPACING FROM DIPOLE
REFLECTOR	41 3/4 IN.	19 IN.
DIRECTOR #1	36 3/4	8
2	36 5/8	28
3	36 1/2	44
4	36 3/8	60
5	36 1/4	76
6	36 1/8	92
7	36	108
8	35 7/8	124

FIGURE 11

144 MC YAGI — 13 ELEMENTS ON A 24 FOOT BOOM — 16.1 DB FORWARD GAIN



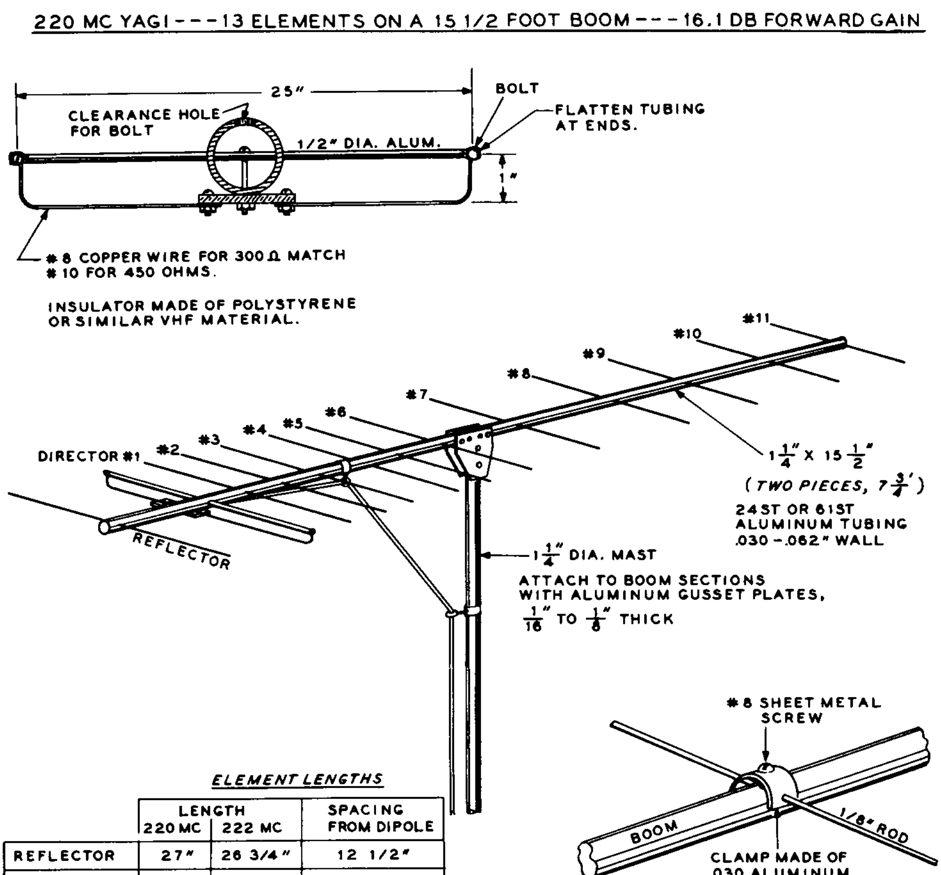




ELEMENT LENGTHS

	LENGTH 144 MC	145 MC	146 MC	147 MC	SPACING FROM DIPOLE
REFLECTOR	41	40 3/4	40 7/16	40 3/16	19
DIRECTOR #1	36 3/4	36 1/2	36 3/16	36 5/16	7
2					14.5
3					22
4					38
5					70
6					102
7					134
8					166
9					198
10					230
11					262

FIGURE 12



ELEMENT LENGTHS

	LENGTH 220 MC	222 MC	SPACING FROM DIPOLE
REFLECTOR	27"	26 3/4"	12 1/2"
DIRECTOR #1	24	23 3/4	4 9/16
2			9 1/2
3			14 7/16
4			24 15/16
5			45 15/16
6			66 15/16
7			87 15/16
8			108 15/16
9			129 15/16
10			150 15/16
11			171 15/16

Fig. 13 All-metal 220 mc Yagi antenna uses eleven directors, all of the same length, providing a maximum gain figure of 16db over the entire band.

Proper stacking distance for this antenna is 16 to 18 feet. If a coaxial feedline is used, a balun transformer should be mounted on the mast.

All parasitic elements are made of 1/8" diameter rod, preferably silver plated steel, or brass. 24ST aluminum is satisfactory, but it should be treated with a thin coating of protective lacquer after assembly to retard oxidation.

220 Mc Yagi—13 Elements on a 15½ Foot Boom—16.1 DB Gain

The 220 Mc Yagi of Figure 13 is very similar to the 144 mc, 13 element

432 MC. YAGI
13 ELEMENTS ON AN 8 FOOT BOOM
16.1 DB GAIN

	ELEMENT LENGTHS			
	LENGTHS 420 MC.	432 MC.	450 MC.	SPACING FROM DIPOLE
REFLECTOR	13 3/4"	13 3/8"	12 7/8"	6 1/4"
DIPOLE	13 1/8	12 3/4	12 1/4	—
DIRECTOR #1	12 5/16	12	11 9/16	2 11/32
2				4 27/32
3				7 11/32
4				12 5/8
5				23 1/4
6				33 7/8
7				44 1/2
8				55 1/8
9				65 3/4
10				76 3/8
11				87

Fig. 14 420 mc long Yagi antenna requires eight foot boom, and produces 16 db gain figure. All directors are same length in this design, allowing maximum gain figure to be achieved.

model. It is built on a 1¼″ diameter boom, but due to the shorter array length it does not require diagonal braces. A boom diameter of 1″ may be used if desired, although braces will then be necessary, and all element lengths must be reduced ⅛″ to allow for the decreased diameter of the boom. Proper stacking distance for this antenna is 10 to 12 feet.

All parasitic elements are made of ⅛″ diamter rod, similar to the 144 mc array.

432 Mc—13 Elements on an 8 Foot Boom—16.1 DB Gain

A scaled down version of the 220 mc antenna will prove highly effective on 432 mc. The boom can be made of ½″ diameter 24ST or 61ST aluminum tubing, eight feet long. The chart of Figure 14 lists element lengths for the ½″ boom. All parasitic elements are made of ⅛″ diameter rod, similar to the 144 mc array.

The folded dipole assembly is made of #8 and #20 copper wire. The #8 wire is passed through the boom and secured with either an external clamp or a sheet metal screw threaded through the boom so as to wedge against the wire, as shown in Figure 2. The #20 wire is supported by the bolts in the insulator at the center feed point, and soldered at the outer tips to the #8 wire. Center-to-center spacing of the two wires should be adjusted to about ⅜″. The #20 wire will provide a 300 ohm termination. A #22 wire will match 450 ohms. Silver plating of the dipole is recommended, otherwise apply a coating of lacquer as explained for the 144 mc array. Proper stacking distance for this antenna is 5 to 6 feet.

12 Element Colinear-Broadside Array—10.5 DB Gain

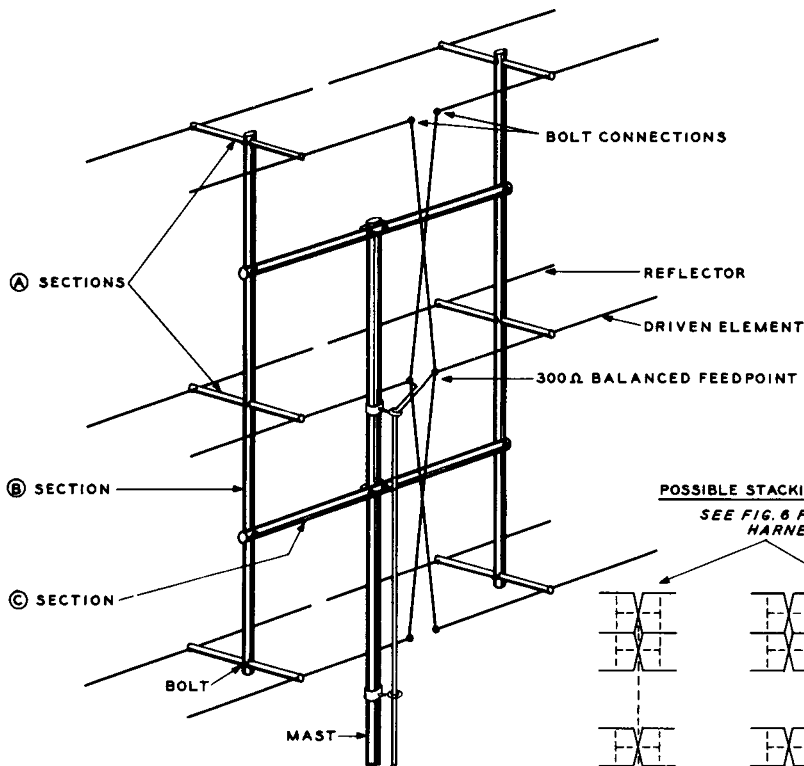
The colinear-broadside array continues to be a popular VHF antenna, in spite of the upsurge in the popularity of long Yagi configurations. Shown in Figure 15 is design information for a 12 element "bed spring" antenna suitable for operation on any one of the VHF bands.

Vertical support members of the frame may be made of metal, but it is

12 ELEMENT COLINEAR – BROADSIDE ARRAY

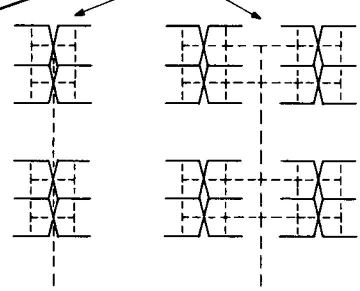
DIMENSIONS FOR 50, 144, 220, AND 432 MC.

POWER GAIN: 10.5 DB OVER A HALF-WAVE DIPOLE



POSSIBLE STACKING COMBINATIONS

SEE FIG. 6 FOR MATCHING HARNESS DATA



DIMENSIONS

	DIAMETER–LENGTH 50 MC.	144 MC.	220 MC.	432 MC.
REFLECTORS	3/8 X 117″	1/4 X 40″	1/4 X 26 1/2″	1/8 X 13 1/2″
DRIVEN ELEMENTS	3/8 X 110″	1/4 X 38 1/2″	1/4 X 24 1/2″	② #8 X 12 3/4″
REFLECTOR SPACING	34″	11 3/4″	7 5/8″	3 7/8″
PHASING LINES ①	114″	39 1/2″	25 3/4″	13 1/8″
Ⓐ SECTIONS	1 X 35″	1/2 X 12 1/4″	1/2 X 8 1/8″	3/8 X 4 3/8″
Ⓑ SECTIONS ③	(2 *PIECES*) 1 1/4 X 230″	3/4 X 81″	3/4 X 52 1/2″	1/2 X 27″
CENTER-TO-CENTER SPACING BETWEEN Ⓑ SEC'TS.	118″	41″	27″	14″
Ⓒ SECTIONS (*WOOD*)	2 X 2 X 121″	1 X 2 X 43″	1 X 2 X 29″	1 X 1 X 15″
MAST SECTION (*MINIMUM*)	2″ DIA.	1 1/4″	1 1/4″	1″

① *PHASING LINES MAY BE MADE OF OPEN WIRE LINE OR TUBING. SPACING IS NOT CRITICAL. THEY MUST BE TRANSPOSED AS ILLUSTRATED.*

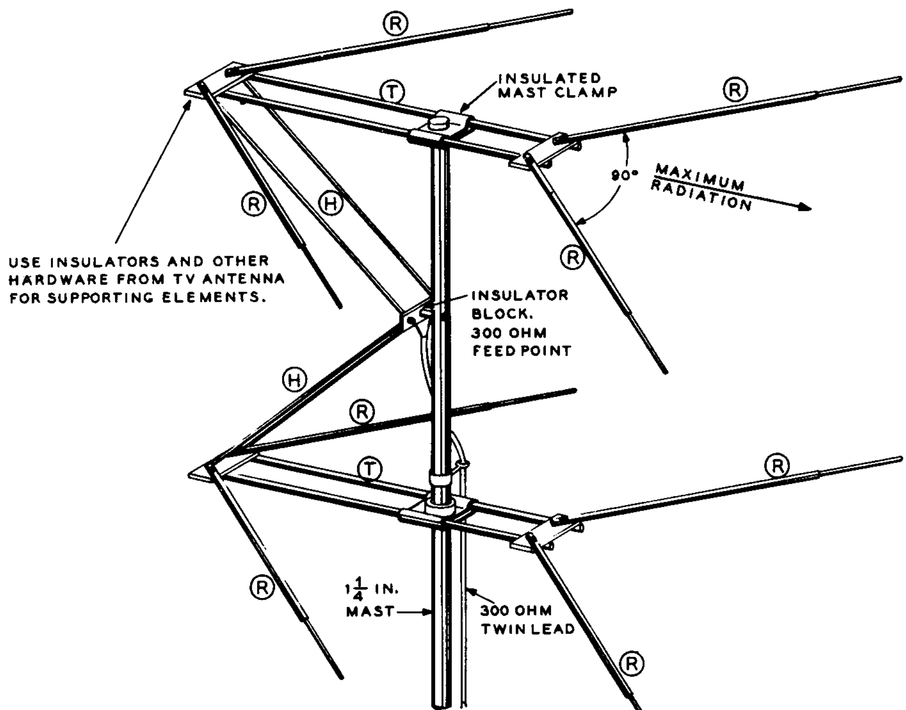
② *COPPER WIRE. SOLDER PHASING LINES TO ELEMENTS.*

③ *THE THREE HOLES IN THE Ⓑ SECTIONS WHICH PASS THE Ⓐ SECTIONS SHOULD BE SPACED TO MATCH THE PHASING LINES.*

FIGURE 15

MULTI-BAND BEAM FOR 50, 144, AND 220 MC.

FIGURE 16



DIMENSIONS

R – 67 IN. (*ORIGINAL TV LENGTH: 50 IN.*)
3/8 IN. TUBING WITH 1/4 IN. DIA. EXTENSIONS.

T – 62 IN. (*ORIGINAL TV LENGTH: 47 IN.*)
1/2 IN. TUBING SPACED 3 1/8 IN. CENTER-TO-CENTER. USE NEW TUBING, OR SPLICE ADDITIONAL SECTION TO ORIGINAL PIECE.

H – 57 IN. 1/4 IN. TUBING, SPACED 4 IN. CENTER-TO-CENTER.

APPROX. FORWARD GAIN OVER HALF-WAVE DIPOLE		FRONT-TO-BACK RATIO	SIDE LOBES
50 MC.	7 DB	10 DB	NIL
144 MC.	12 DB	18 DB	-12 DB
220 MC.	13.5 DB	18 (*EST.*)	-8 DB

necessary that the horizontal sections (C) be made of wood, as they lie in the electric plane of the elements. The complete structure may be made of wood, if desired.

The array is fed at the center point with a 300 ohm balanced transmission line. If coaxial feed is desired, a balun should be mounted on the mast a few feet under the array.

This basic antenna may be stacked in various combinations, using the matching harness assembly illustrated in Figure 6.

Multi-band Beam for 50, 144, and 220 Mc.

The following description shows how a familiar TV antenna design may be modified to serve as a multi-band VHF beam for use in the 50, 144, and 220 mc amateur bands. As a matter of fact, the antenna will also perform well in the 420 mc band if open wire feedline is used.

Electrically, each bay of the antenna is a pair of half wave dipoles at 50 mc, one of them spaced ¼-wavelength in front of the other. They are

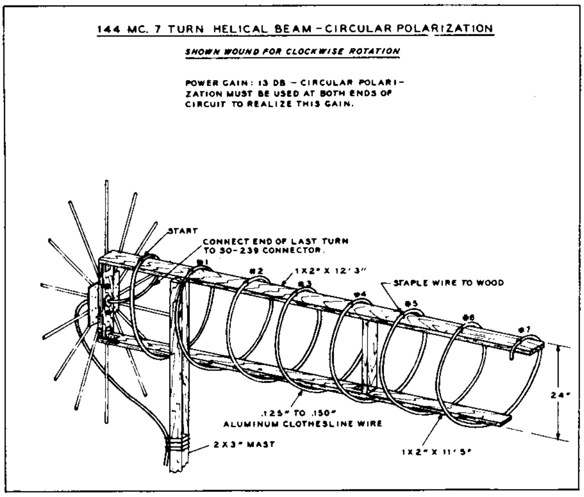


Fig. 17 Helical antenna provides 13 db power gain over 2:1 frequency range. The same direction of rotation of the coil must be employed at both ends of the circuit to achieve maximum gain figure.

driven with a phase difference of 90 degrees by means of the horizontal phasing line (T). This feed method produces a unidirectional radiation pattern having a good F/B ratio. At 144 mc the antenna functions on its third harmonic, and begins to operate as a short V-beam, providing a little more gain than at 50 mc. At 220 mc the V-beam is working at the 5th harmonic. Naturally, the apex angle and other dimensions represent a compromise in order to achieve broad band operation.

Two of these bays are stacked to provide an additional 3 db of gain, and are fed in phase with an open wire transformer (H). The two bays are separated about ½ wavelength at 50 mc, 1½ wavelengths at 144 mc, and 2½ wavelengths at 220 mc.

Spacing between the upper and lower bays is determined by the positioning of the matching transformer (H), and will be about 74 inches. Feedpoint impedance is quite close to 300 ohms on 50 mc and 144 mc, but drops to about 200 ohms at 220 mcs. This is not serious if the twin lead feedline is less than about 75 feet in length. For longer spans, it is best to use open wire line. A suitable antenna coupler for tuning out line reactance will prove handy at 220 mc and 420 mc, although the length of the transmission line may be pruned for best loading at these frequencies. Gain runs 7 db at 50 mc, 12 db at 144 mc, and 13.5 db at 220 mc.

144 Mc Helical Beam—13 DB Gain

Shown in Figure 17 is a simplified design for a 144 mc helical beam antenna. It is wound for clockwise rotation, and the same direction of rotation must be employed at both ends of a circuit to realize the maximum gain figure of 13 db. The frame of the antenna is made of two pieces of 1″ x 2″ wood, spaced about 2 feet apart. The helix is made of a length of aluminum clothesline wire wound about the frame and stapled to the wood. A ground plane is placed at the base of the helix. This plane is made of a foot-square section of aluminum sheet, with 14 projecting radials, each three feet long. A SO-239 coaxial receptacle is placed at the center of the ground plane, and the base of the helix is attached to the center conductor of the receptacle.

The feedpoint impedance of the helical beam is about 130 ohms, so a ¼-wave matching transformer made of a 27″ length of RG-11/U coaxial line must be used between a 52 ohm transmission line and the antenna.

CHAPTER VIII

VHF Receiver Design

Receiver sensitivity is an important factor of receiver design in the VHF region. In many instances it is the key that will spell success or failure of VHF long distance communication. Paradoxically, receiver sensitivity is an entirely different problem in the VHF region than it is in the high frequency portion of the radio spectrum. The noise level external to the receiver at those frequencies that are reflected by the ionosphere is extremely high, thus limiting the useable sensitivity of the receiver. The very solar radiations that create the ionized layers of the earth's atmosphere blast the earth with a continual din of radio noise. In addition, various forms of galactic noise from outer space contribute to the high frequency background noise.

The level of this overall cosmic noise varies from day to day, and from hour to hour, as the emanations from the sun wax and wane. Superimposed upon this natural noise is a variable amount of atmospheric noise caused by the interplay of electrical forces in the earth's atmosphere. These, too, vary to a large degree from time to time.

It can be observed that these radio noises form a floor, or minimum level beneath which reception of radio signals is extremely difficult. Receivers of even mediocre design have sufficient sensitivity to reach the external noise level of the high frequency spectrum. Indeed, in large cities having an abundance of electrical appliances and power generating systems, the man-made noise level is so high that a sensitive high frequency receiver is rendered almost useless.

An approximation of the cosmic noise level as a function of frequency is shown in Figure 1. This random noise is sufficiently strong enough at 50 mc to be heard over the internal noise level of a good receiver. However, as the frequency of observation is increased, the external noise level drops to such a low value that it may be neglected when compared to the internal noise level of the VHF receiver.

The minimum radio signal that may be detected in the VHF region is therefore not so much a function of various types of external noise which cannot be controlled, but more a function of internal receiver noise, which may be minimized by the use of proper design techniques. Special efforts have been made over the past few years to design circuits and tubes that would contribute a minimum amount of noise to the received signal.

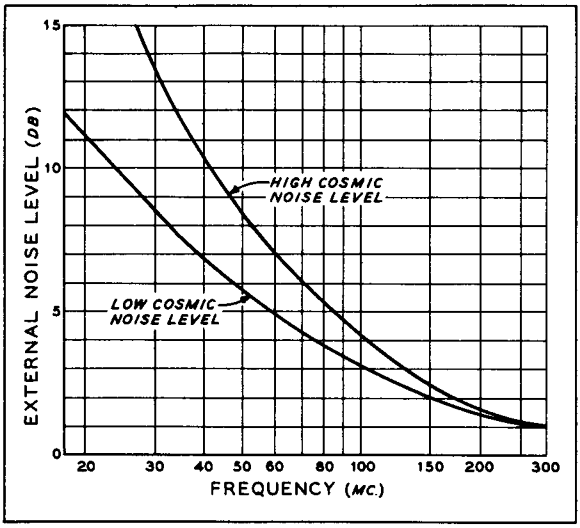


Fig. 1 Approximate cosmic noise level decreases as operating frequency is increased. Receiver noise level becomes limiting factor above 50 mc or so.

VHF Radio Frequency Amplifiers

A general introduction to VHF noise problems has been discussed in chapter five of this Handbook. As stated, the noise figure of the first r-f stage of the receiver will generally determine the overall figure of merit for the whole receiver. Any noise generated in the first stage is of major importance because it receives amplification through the entire receiver. Noise generated in succeeding stages is of less consequence as the signal level becomes sufficiently high to override the noise levels encountered. Thus, the sensitivity capabilities of a receiver are effectively limited by the noise produced in the initial stages. The noise factor, which can be considered as a measure of degradation of the signal-to-noise power ratio as it passes from the antenna through the receiver, provides a convenient method of specifying the noisiness of a receiver. When the minimum useable signal is determined by receiver noise, a reduction in noise factor is equivalent to an actual increase in transmitted power. Therefore, in developing a VHF radio frequency amplifier, primary attention must be directed to achieving a minimum noise factor.

To obtain low-noise operation it is essential that the incoming signal be amplified sufficiently in the initial stage to override noise levels of succeeding stages. In this way, only a minimum of noise is added to the signal by the receiver. To realize this requirement both the input impedance and gain of the first stage should be high. With a high input impedance, a voltage step-up can be realized in the input circuit. The greater voltage that can be developed at the grid of the first tube means less tube noise will be added to the signal. Similarly, the higher the voltage amplification of the radio frequency amplifier, the less will be the noise contributed by the succeeding mixer or amplifier stages. These considerations dictate the value for minimum gain that is acceptable in the r-f stage.

The Pentode R-F Amplifier

The circuit of a typical VHF pentode r-f amplifier is shown in Figure 2. There are several advantages inherent in the pentode amplifier. It is capable of operation with relatively high gain. Neutralization is not required in

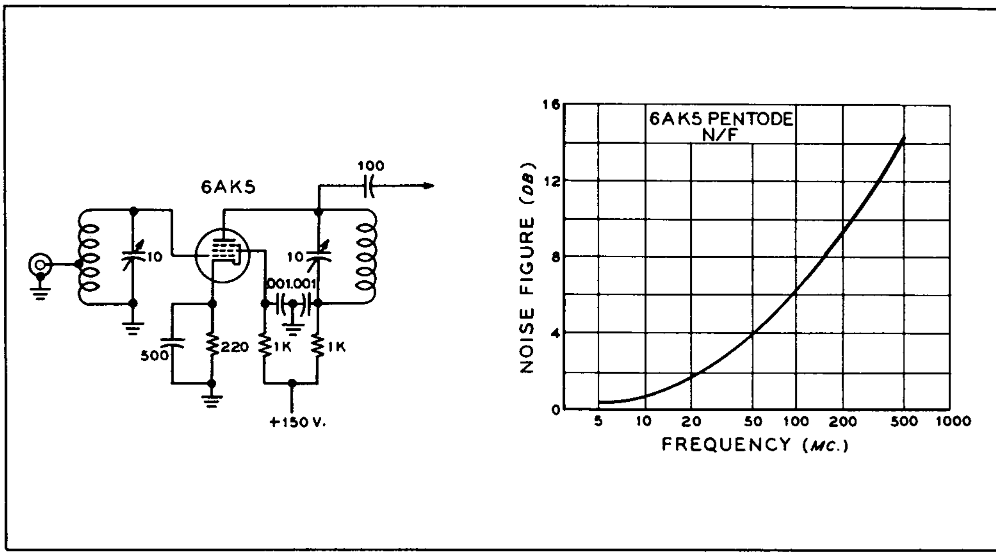


Fig. 2 Noise figure of pentode amplifier increases rapidly above 50 mc. Poor N/F is due to partition of cathode current between screen and plate.

adequately shielded circuits, and the input impedance varies only slightly with automatic gain control bias. Unfortunately, however, the pentode is restricted by its high-noise characteristic. In addition to the shot-effect noise, common to both triodes and pentodes, there is partition noise which results from irregularities in the division of the cathode current between the screen grid and the plate. As a result, the noise of a pentode is from three to five times that of the same tube connected and operated as a triode.

Certain pentodes are capable of operation as high as 400 mc or so, although the noise figure for any pentode deteriorates rapidly with increasing frequency, as shown in Figure 2.

The Grounded Grid R-F Amplifier

A typical grounded grid r-f amplifier is shown in Figure 3. Although the grounded grid triode can offer improved noise characteristics over the pentode, the tube and circuit suffer from low gain and low input impedance. It is usually necessary to employ two grounded grid stages to achieve a suitable gain figure. The upper frequency limit of a grounded grid amplifier employing conventional VHF tubes is in the region of 800 mc or so, at which point the tube efficiency drops because of transit time loading effects.

In the circuit of Figure 3, the grid of the tube is at ground potential, and acts as a shield between the input and output circuits. The input impedance is in the neighborhood of several hundred ohms, and precautions must be taken to prevent excessive circuit loading by this low value of impedance. In most cases a satisfactory impedance match may be obtained by tapping the cathode of the tube down the input circuit until a good impedance match is obtained. Since the cathode is capacitively coupled to the filament circuit it is necessary to employ nonresonant filament chokes to reduce the shunting effect of the filament upon the input circuit.

Tubes which lend themselves well to the grounded grid circuit include the 6BC4, 6J4, 6AJ4, and various versions of the disc-seal tubes. Shown in Figure 4 are the GL-6BY4 and GL-6442 VHF tubes designed for service as

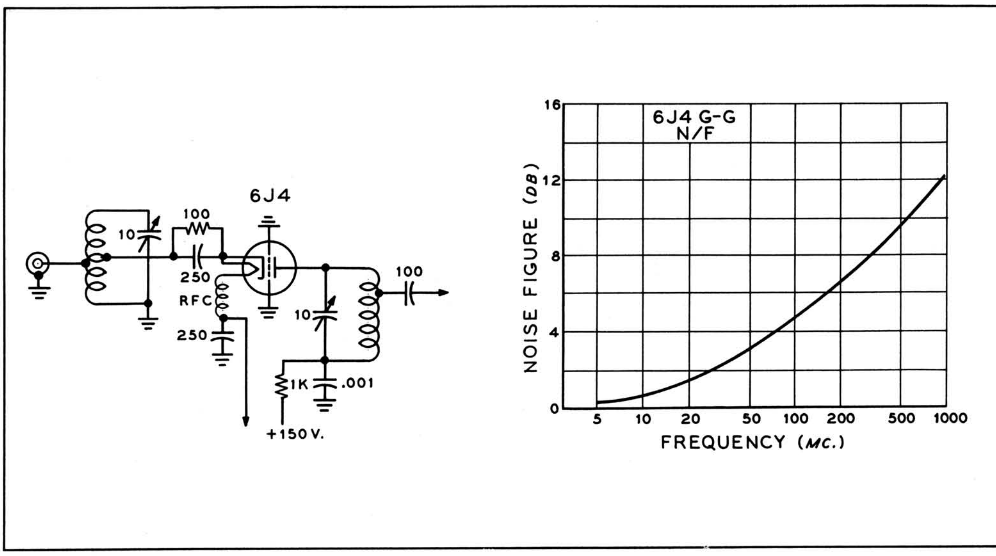


Fig. 3 Noise figure of g-g amplifier is better than that of pentode, but circuit suffers from low stage gain and relatively low input impedance.

grounded grid amplifiers for transmission and reception. The planar construction using very close element spacings reduces transit time effects, and the large grid flange permits a low inductance path between the grid element and ground. The GL-6BY4 will provide a noise figure of better than 8.5 db at 900 mc when the tube is used in a grounded grid coaxial tank circuit.

The Neutralized Triode Amplifier

Triode r-f amplifier stages of various types have become quite popular for VHF reception. A typical grounded cathode push-pull triode r-f amplifier is shown in Figure 5. This circuit provides a higher gain figure than the grounded grid arrangement. Because of the relatively large grid-plate capacitance of the triode tube, it cannot be employed as an amplifier in conventional circuits unless neutralization is employed. The neutralization adjustment is relatively uncritical, and for work at 50 mc and 144 mc, the push-pull 6J6 amplifier has proven to be most effective, producing an extremely low noise figure, yet exhibiting a high degree of stability.

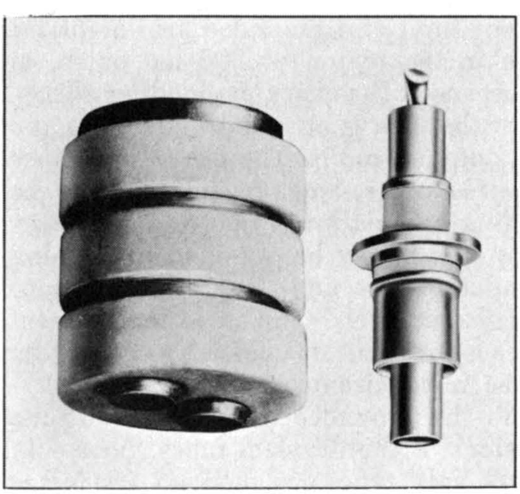


Fig. 4 New planar tubes are designed for VHF service. GL-6442 (right) is only 2¼" long, but will provide over 1 kw peak power. Miniature GL-6BY4 (left) is only ⅝" high. It is first of new ceramic tubes.

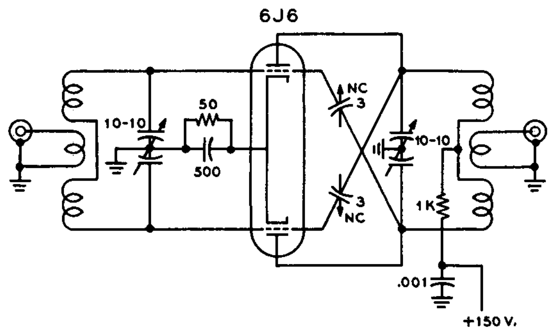


Fig. 5 Push-pull triode amplifier will show a good noise figure in the 144 mc band. Tube must be neutralized to stop oscillation and to allow highest noise figure.

The Cascode R-F Amplifier

The above discussion indicates the need for an improved tube and circuit combination to perform as an r-f amplifier in the VHF region. The inescapable noise characteristics of pentode tubes necessarily indicate that triode amplifiers must be employed for best low noise results. In view of the previously mentioned gain requirements, and assuming the order of transconductance now obtainable from modern triode construction, two stages of triode amplification seem to be in order.

These two stages can be arranged in a variety of ways. Theoretical and experimental investigations indicate that the combination of a grounded cathode stage followed by a grounded grid stage yields optimum performance. It can be shown that this combination produces the high amplification and stability of a pentode, yet has the desired noise characteristics of a triode.

The grounded cathode amplifier followed by a grounded grid amplifier

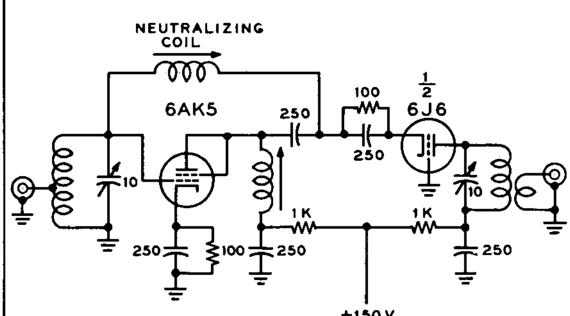


Fig. 6 Cascode amplifier combines good N/F of a triode amplifier with the high gain of a pentode stage. First tube of the cascode must be neutralized for best noise figure.

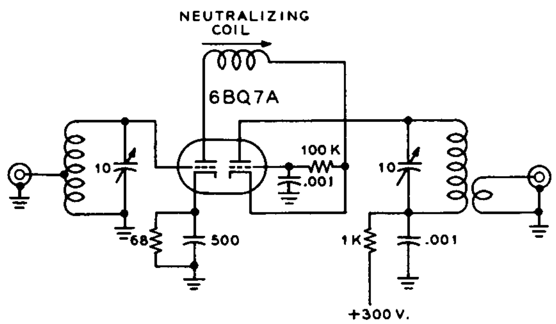


Fig. 7 Special triodes have been designed for the cascode circuit. At left is series cascode, with same plate current flowing through both sections of the tube.

was developed at the M.I.T. Radiation Laboratories, by H. Wallman and provides an excellent noise figure. The term *cascode* is now generally used to describe such a circuit. A typical cascode amplifier is shown in Figure 6. For best noise figure, the first triode amplifier should be neutralized. This may be accomplished inductively, as shown in the drawing.

The overall noise figure of the cascode amplifier depends upon the choice of input tube. A high transconductance tube having low interelectrode capacitances should be chosen. Some of the better tubes for this service are the 6J4, 6BY4, 6AJ4, and 6BC4. Certain double triodes, such as the 6BQ7-A, 6BK7, and 6BZ7 are especially designed for VHF cascode service. Each triode section of these tubes has a separate cathode, and the two units are electrically independent. To minimize interstage coupling an internal shield is located between the two triode sections.

When the double triodes are employed as a series cascode amplifier, as shown in Figure 7, the same plate current flows through each section. In addition, the cathode of the grounded grid section operates at a relatively high positive d-c potential with respect to the heater. The cascode-type triodes have a 250 volt heater-cathode rating permitting this type of service. Other double triodes should employ the parallel cascode circuit, shown in Figure 6 which removes the high potential from the cathode of the second triode. Operation of the two circuits is identical.

The amplification of a cascode amplifier, like that of a pentode, is proportional to the transconductance of the tubes. The amplification of the first stage may be expressed as a ratio of the transconductances of the two tubes, and usually runs between 0.5 and 1.25. The overall circuit gain may run as high as 300 in the lower portion of the VHF region.

The Neutrode VHF Amplifier

The Neutrode circuit is a modern adaptation of the old Neutrodyne amplifier, popular in the days of the WD-11, UV-199 and UX-201A tubes. Shown in Figure 8A is a plate neutralized triode amplifier based upon the Neutrode circuit. The equivalent bridge neutralizing circuit is shown in Figure 8B. Best noise figure and highest amplification are obtained when

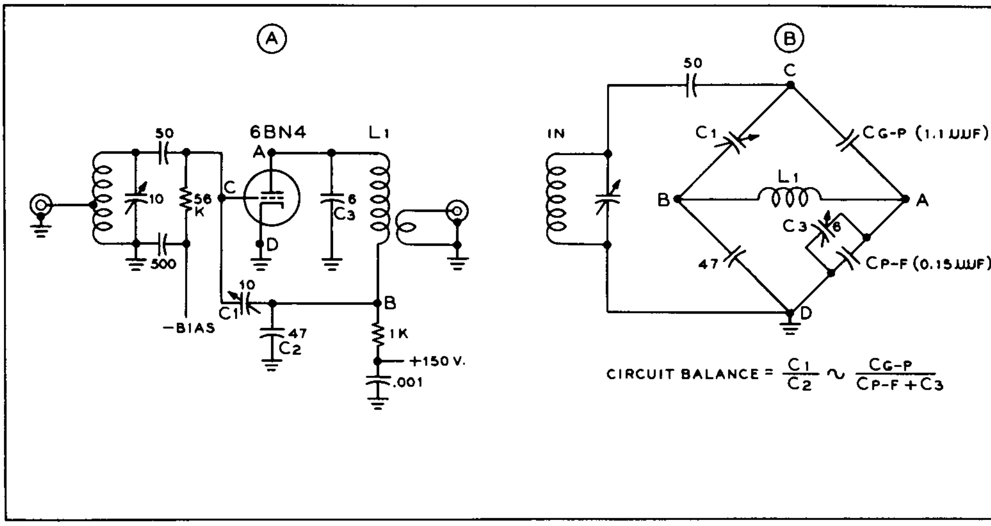


Fig. 8 Plate neutralized triode amplifier (Neutrode) is employed as a low noise, high gain amplifier in inexpensive television tuners.

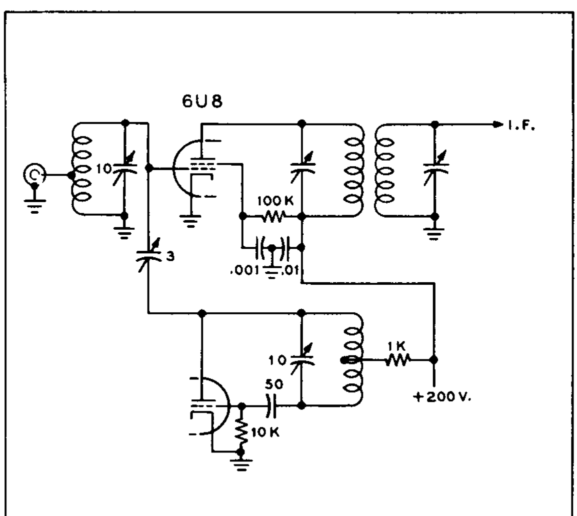
the capacity bridge is balanced by adjustment of the neutralizing capacitor, C1. Plate voltage is removed from the amplifier for this adjustment, and C1 is tuned for minimum signal leakage through the stage. Final neutralizing adjustment should be made with a noise generator to achieve the best possible noise figure.

The 6BN4 triode has dual cathode and grid leads to reduce the residual inductance in the neutralizing circuit to a minimum. A voltage gain of about 40 may be obtained with this amplifier, with a noise figure closely approaching that of the dual triode cascode circuit. The simplicity of the Neutrode amplifier as compared to the cascode amplifier makes it popular in inexpensive television tuners and simple VHF amplifiers.

Frequency Conversion

Frequency conversion, at best, is a noisy process. Converter tubes, generally speaking, have a much higher noise level than any r-f amplifier tube

Fig. 9 Converter circuit using dual purpose 6U8 tube is suitable for use at VHF if low noise r-f stage is placed ahead of it.



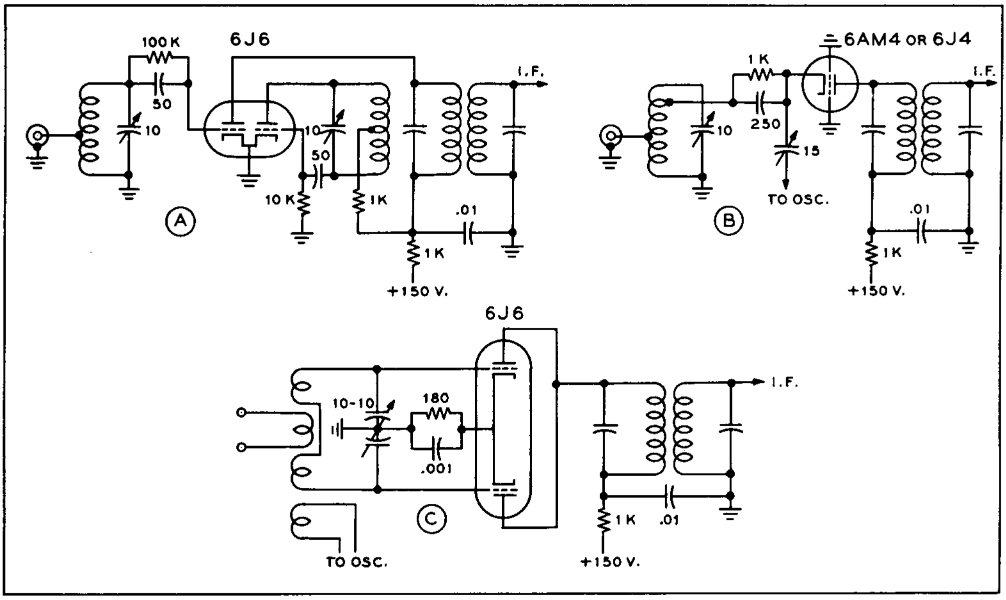


Fig. 10 Low noise triode mixers are effective in VHF region. Grounded grid mixer (B) performs well into the UHF region when used with new planar tubes.

and exhibit a transconductance that runs about one-third the transconductance obtainable when the tube is operated as an amplifier. Multi-element converter tubes, such as the 6BA7 and 6BE6 are especially noisy, and should be avoided wherever possible since they require an extremely high input signal to overcome the electron noise generated by the mixing process.

Certain converter tubes, such as the 6U8 perform in a satisfactory manner in the VHF region. The pentode mixer offers high conversion transconductance, and high input and output impedances. A typical triode-pentode converter circuit using this tube is shown in Figure 9. The noise figure of such a circuit is of the order of 14 db at 144 mc.

Triode mixers have been employed with good results in the VHF region. The noise figure of such a mixer is measurably lower than that of a pentode mixer, but not as good as when the same triode is employed as an r-f amplifier. A typical triode mixer is shown in Figure 10A.

Tests have shown that a much lower level of oscillator injection can be used for high-mu mixer tubes than for low-mu tubes to give the same value of noise figure. Suitable high-mu mixer tubes for VHF work are the 6J6, 12AT7 (and its cascode counterparts), 6J4 and 6AM4. The latter two tubes may be employed in a grounded-grid mixer configuration, as shown in Figure 10B.

A third form of mixer that is widely employed for VHF work is the push-pull circuit, as shown in Figure 10C. This circuit performs well at 144 mc and 220 mc when a 6J6 dual triode tube is used.

The Conversion Oscillator

The conversion oscillator furnishes an output from the converter stage that differs from the signal frequency by the value of the intermediate frequency. In the lower portion of the VHF spectrum, various triode tubes can be employed as self-excited oscillators, using mutually coupled circuits

to maintain feedback. As the UHF region is approached the use of mutually coupled coils becomes difficult, and resonant lines and cavities take their place.

Stability of a self-controlled oscillator is always a problem in the VHF range. A regulated plate supply should be employed, and the oscillator components should be ruggedly built and mechanically stable. The oscillator tuning capacitor should have two bearings, and the plates should be heavy, and preferably double-spaced. All leads should be made of heavy, solid copper wire, and the oscillator coil should be mechanically rigid.

When maximum frequency stability is desired, it is common to employ a crystal controlled conversion oscillator. If it is desired to cover a small tuning range, the intermediate frequency amplifier may be tuned. The most common arrangement for obtaining a high frequency conversion signal is to employ a low frequency crystal oscillator, followed by several doubler stages.

The Intermediate Frequency Amplifier

The intermediate frequency amplifier of any receiver is employed to build up the strength of the received signal, and to restrict the bandwidth of the receiver. A common rule of thumb is that the intermediate frequency should not be more than 10% of the operating frequency. This allows a balance to be reached between i-f selectivity and image response. Thus, for 144 mc operation, the intermediate frequency should be in the region of 15 mc. If greater selectivity is desired than is obtained from this choice of i-f, the best solution is to employ a second conversion to a lower intermediate frequency. For example, the 15 mc intermediate frequency could be converted to 455 kc by the use of a crystal controlled converter stage.

The signal-to-noise ratio of the receiving system improves as the passband of the receiver is decreased. For reception of scatter and moon-bounce signals, it is mandatory that i-f bandwidths of less than one kilocycle be employed. Noise may be received over a relatively large portion of the spectrum with a receiver having a wide i-f passband. When the "door" to the radio spectrum is partially closed by narrowing the i-f passband, a proportionately smaller amount of external noise will reach the second detector of the receiver. If the passband of the receiver is just wide enough to admit the desired signal, a minimum of noise will pass through the "door" into the receiver circuits. A minimum passband of about 3 kc (measured at the 3 db points of the i-f response curve) is required for voice reproduction, and a passband of 800 cycles or less may be employed for c-w reception.

CHAPTER IX

VHF Receiver Construction

The construction of VHF equipment is a fascinating art. The necessary small components and midget tuned circuits can be combined to make a true masterpiece. By proper design, items of VHF equipment do not require a machine shop and a collection of many tools to permit fabrication and assembly. Indeed, many of the units shown in this chapter may easily be constructed with a minimum of tools, using the kitchen table for a workbench.

VHF Converters

The best VHF receiver takes the form of a highly stable and sensitive converter working into a good low frequency receiver. Many advantages are realized when this type of reception is employed. The high frequency conversion oscillator may be crystal controlled, giving the receiver an excellent

COMPONENT NOMENCLATURE
FOR SCHEMATICS

CAPACITORS

1. VALUES BELOW 999 μμF ARE INDICATED IN UNITS.
   EXAMPLE: 250 μμF DESIGNATED AS 250
2. VALUES ABOVE 999 μμF ARE IND. IN DECIMALS.
   EXAMPLE: .0015 μF DESIGNATED AS .0015
3. ALL OTHER CAPACITOR VALUES ARE AS STATED.
   EXAMPLE: 15 μF, 0.3 μμF, ETC.
4. TYPE OF CAPACITOR IS DESIGNATED AS FOLLOWS:
   M = MICA
   BM = BUTTON MICA
   C = CERAMIC
5. VOLTAGE RATING OF CAPACITORS IS 600 VOLTS UNLESS OTHERWISE STATED.

RESISTORS

1. ALL RESISTORS 0.5 WATT COMPOSITION TYPE UNLESS OTHERWISE NOTED.
2. RESISTANCE VALUES STATED IN OHMS, THOUSAND OHMS (K), AND MEGOHMS (M)

EXAMPLE:
330 OHMS IS DESIGNATED 330
1500 OHMS IS DESIGNATED 1.5 K
1,000,000 OHMS IS DESIGNATED 1 M

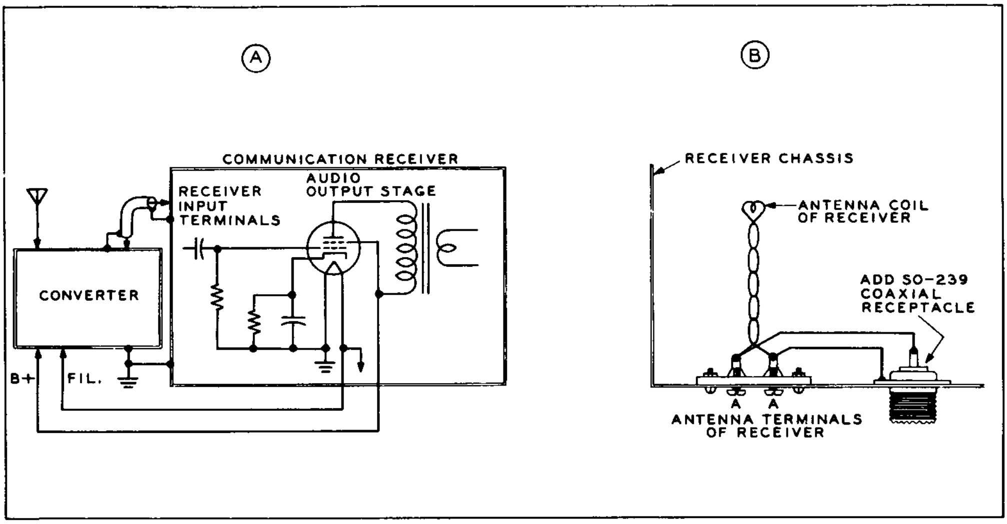


Fig. 1 Correct method of connecting VHF converter to receiver.

order of stability. The first intermediate frequency should be in the neighborhood of 7 to 30 mc (depending upon the operating frequency of the converter) permitting a high order of rejection of signals out of the normal passband of the converter. Finally, the cost of such a converter is a small item when compared to the cost of building a complete VHF receiver. Because of these reasons, the crystal controlled converter is by far the most popular device for VHF reception. Readers interested in a simple, tunable VHF converter are referred to the *Novice and Technician Handbook*, published by Radio Publications, Danbury Rd., Wilton, Conn. Many simple items of VHF equipment for the beginner are described in this Handbook.

The Converter and the Receiver

Attention should be paid to the method of connecting the VHF converter to the receiver. In some instances, power to operate the converter may be "stolen" from the audio tube socket of the receiver, as shown in Figure 1A. Converters employing four or more tubes usually impose too great an additional burden upon the power supply of the receiver, and therefore require a small power unit of their own.

It is necessary to take steps to prevent the station receiver from picking up spurious signals when it is tuned over the intermediate frequency range of the converter. Many powerful shortwave stations operate in the 7 mc to 30 mc range, a portion of which is usually employed as the i-f tuning range of the converter. If these signals are allowed to reach the input circuits of the station receiver, they will ride through the receiver, showing up as interfering signals within the tuning range of the VHF amateur band. It is therefore necessary to employ a shielded coaxial line between the converter and the station receiver to reduce spurious i-f pickup. The shield of the coaxial line should be well grounded to the chassis of the units at each end. If possible, the chassis of the VHF converter should be physically attached to the chassis of the station receiver to insure that both units are at the same i-f ground potential. Coaxial connectors should be employed at both ends of the connecting i-f cable for best i-f spurious signal suppression. A coaxial receptacle may be mounted on the rear apron of the station receiver

(Figure 1B) and connected in parallel with the antenna terminals of the receiver if one is not already provided.

Converter Adjustment

The procedure to be followed in converter adjustment and alignment is essentially the same, regardless of the design of the unit. The first step is to check all the wiring for errors before power is applied to the converter. If a grid dip oscillator is handy, the resonant frequency of each tuned circuit should be measured with all the converter tubes in their sockets. Slight adjustments should be made to the coils to bring the circuits to the approximate resonant frequency.

The second step is to adjust the crystal oscillator and doubler stages. A #48 or #49 (2-volt, 60 ma) pilot lamp with a 1-inch diameter loop of wire soldered to the base terminals is held near the oscillator coil, and the tuning capacitor adjusted until the lamp glows. A sluggish crystal may require a turn or two added to the feedback coil to sustain oscillation. Excessive feedback will be noticed as variations in lamp brilliancy, or as "birdies" heard in the station receiver when the latter is attached to the converter.

The final step is to adjust the coupling between the various stages of the converter for optimum noise figure, as discussed in chapter 12. The first adjustment to make is to the antenna coupling circuit. The antenna tap on the r-f coil, or the antenna link coil should be varied for best N/F. If a cascode amplifier is used, the filament circuit should be opened, and the neutralizing coil adjusted for minimum signal leakage when a signal is applied to the antenna circuit. The second step is to adjust the coupling between the r-f amplifier and the following stage. Too much coupling will result in spurious signals, while too little coupling will result in a deterioration of the noise figure. Coupling capacitors should be tapped along the inductances to obtain the best value of N/F.

The final step is to adjust the level of oscillator voltage injection. This may be done by varying the oscillator-mixer coupling capacitor, or by changing the degree of coupling between the injection coils, if inductive injection is employed. In some instances the N/F may be materially improved by hand selecting the tube employed in the first r-f stage.

The final check is to measure all electrode voltages of the tubes to make sure that they are being operated within the specifications set by the manufacturer.

A Sensitive and Stable Converter for the Beginner

A simple and highly sensitive converter for 50 mc and 144 mc is shown in Figures 2, 3 and 4. Employing only three tubes, this unit is easy to build and requires no extensive tuning adjustments. On 50 mc, a N/F better than 5 db may be obtained, and on 144 mc a N/F better than 7 db may be obtained. The converter requires 250 volts at 30 ma for the plate supply, and 6.3 volts at 0.85 amperes for the filament supply. This amount of power may safely be drawn from the supplies of most of the larger short wave receivers. If the converter is to be used with a small receiver of the ac-dc type, it will be necessary to construct a power supply having the aforementioned capability.

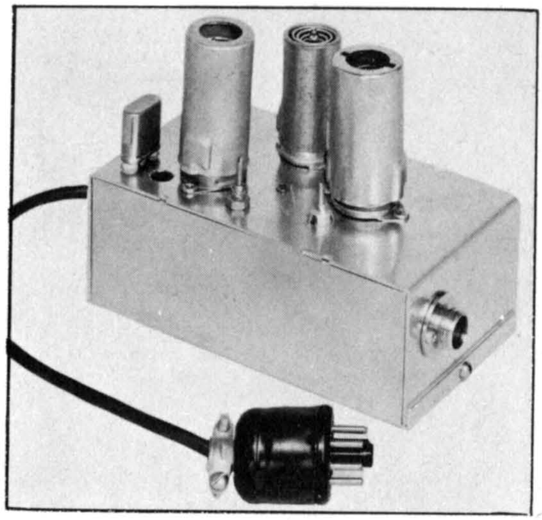


Fig. 2 Three tube 144 mc converter has N/F better than 7 decibels. 12AT7 crystal oscillator is at left and 6BQ7A r-f amplifier-mixer is at the right, with 6AB4 cathode follower to the rear. The oscillator tuning capacitor (C4) may be adjusted through chassis hole to right of the crystal socket. Antenna receptacle is at right end of the chassis.

Converter Circuit

The schematic of the three tube converter is illustrated in Figure 3. One section of a 6BQ7A VHF double triode is employed as grounded grid r-f amplifier. The input circuit is composed of coil L1 and the cathode-ground capacitance of the tube. This circuit is broadly resonant over a range of five or six megacycles. The plate circuit of the low noise r-f amplifier is resonated to frequency by capacitor C1, plus the combined output capacitance of the r-f stage and the input capacitance of the first mixer. This high-C circuit provides the greater part of the r-f selectivity of the converter.

The second section of the 6BQ7A serves as a triode mixer to the intermediate frequency range of 14-18 mc. The output of the mixer is capacitively coupled to a 6AB4 high-mu triode which serves as a cathode follower, providing a low impedance untuned output circuit suitable for matching the input impedance of the station receiver to the mixer circuit of the converter.

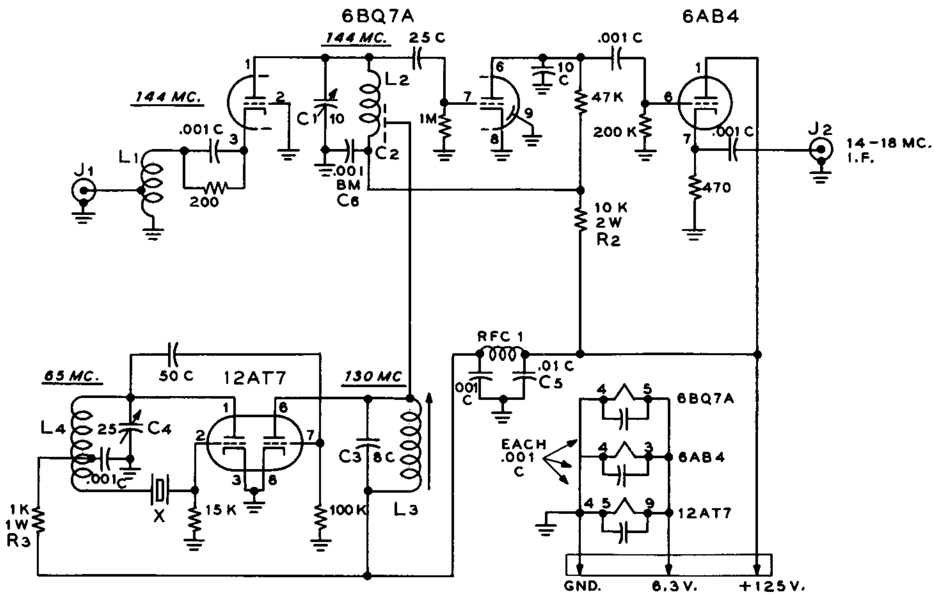
A conversion frequency of 130 mc is needed for 2 meter operation, and is obtained from a 65 mc overtone crystal oscillator and a triode doubler stage. These two stages are contained within a single 12AT7 double triode. The conversion frequency is capacitively coupled to the input circuit of the mixer by a small capacitance (C2) made of a short length of wire.

The circuit of the low noise r-f stage has been considerably simplified by the omission of filament r-f chokes, usually employed in "hot-cathode" grounded grid circuits. Removal of the filament chokes, however, does not alter the noise figure of the amplifier stage and drops the gain figure but slightly. They are therefore omitted for simplicity's sake.

Converter Assembly

The low noise converter is built completely upon one side of a small aluminum box-chassis measuring 5¼"x3"x2⅛" (*L.M.B. #136*). Placement of the major components may be seen in the top and bottom photographs. A chassis drilling template is given in Figure 5.

After the major components have been mounted, the various pins of each tube socket should be grounded as indicated in Figure 3 by means of soldering lugs placed beneath the socket retention nuts. The center stud of



C1 - 10 μμF TRIMMER. *ERIE 532-10*

C2 - APPROX. 2 μμF. SEE TEXT.

C3 - 8 μμF CERAMIC

C4 - 25 μμF VARIABLE CERAMIC. *CENTRALAB 822AZ*

L1 - 4 TURNS # 16E., 3/8" I.D., 1/2" LONG.

L2 - 2 3/4 TURNS # 16, 3/8" I.D., 3/8" LONG.

L3 - 4 TURN # 16E., 1/4" D., 3/8" LONG ON *J.W. MILLER 4400* SLUG TUNED FORM.

L4 - 5 TURNS # 16, 3/4" D., 1/2" LONG, *B&W 3010.*

RFC 1 - 10 MICROHENRIES. *J.W. MILLER 4612.*

X - 65 MC. OVERTONE CRYSTAL.

Fig. 3 Schematic, beginner's 144 mc converter.

each socket should also be grounded, and a short, heavy ground lead is run from pin 2 of the 6BQ7A socket to the mounting stud of C1. The filament leads and filament bypass capacitors should next be wired, followed by all resistors and capacitors that mount between the pins of the tube sockets. The coils should be wound from the data of Figure 3, and soldered in position, and all high voltage wiring completed. As a final step, the three wire power cord is now attached to the converter. The ground lead of the cable should be attached to pin 4 of the 6AB4 socket, and the filament lead should connect to pin 3 of the same socket. The B-plus lead of the power cable should attach to a phenolic tie-point mounted between the 6AB4 and 12AT7 sockets. The oscillator plate circuit decoupling resistor, R3, attaches to this tie-point, as does bypass capacitor C5. The plate bypass capacitor of the r-f stage, C6, is a chassis mounting button mica unit, such as the *Centralab type ZA.*

The input coil, L1, is supported by its leads between the resistor-capacitor assembly attached to pin 3 of the 6BQ7A socket, and the grounding clip of plate tuning capacitor C1. Care should be taken to insure that coils L1 and L2 are at right angles to each other. Coil L2 is supported at one end by the mounting terminal of C1, and at the other end by the center terminal of C6, which is placed halfway between the two nine pin tube sockets.

The doubler coil, L3, is wound upon a slug-tuned coil form which is mounted to the chassis near the 12AT7 tube socket. The oscillator coil, L4, is air-wound and self-supporting between one pin of the crystal socket and

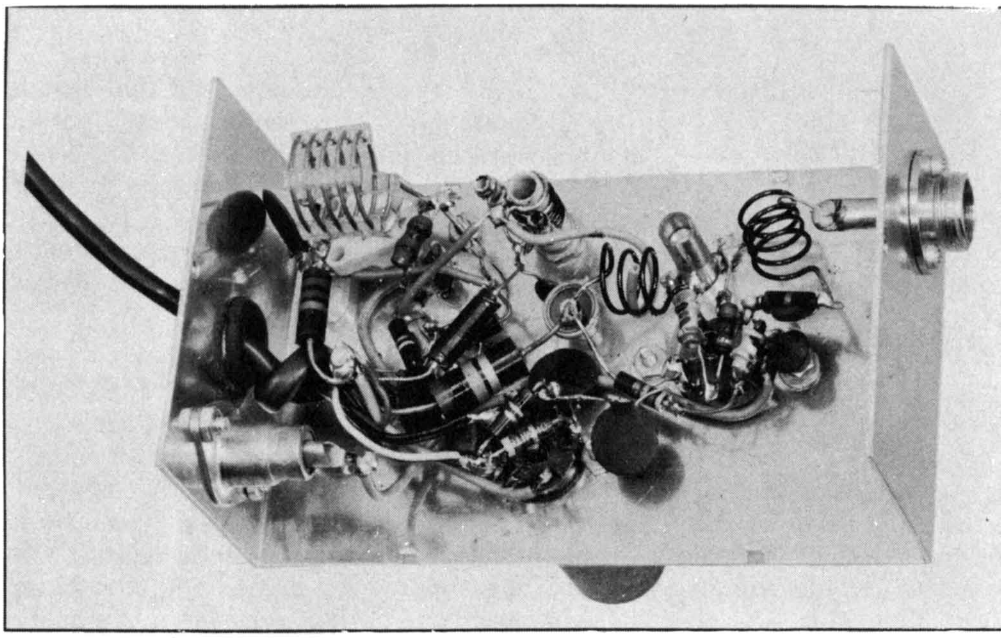


Fig. 4 Under chassis view of 144 mc converter. The 12AT7 socket is at the upper left, in front of L4-C4. L3 is upper center, with C6 in front of it. At the right are the r-f coils, L1 and L2, with C1 behind the 6BQ7 socket. Output recepticle J2 is mounted on rear lip of chassis at left, next to power cable. R2, R3, and RFC-1 are supported by phenolic tie point at left

one terminal of the oscillator tuning capacitor, C4, as seen in Figure 4.

When all wiring has been completed, it should be rechecked for errors before power is applied to the converter.

If the chassis layout and coil design have been copied closely, no trouble should be encountered in obtaining converter operation.

After preliminary adjustments have been made as described earlier in this chapter, a 0-100 d.c. microammeter should be connected to pin 7 of the mixer section of the 6BQ7A through a 100,000 ohm composition ½-watt resistor which may be temporarily soldered to the socket pin. The coupling capacitor C2 should be moved in relation to coil L2 until a reading of 50 to 60 microamperes is obtained on the meter. This insures proper local oscillator injection voltage for the converter.

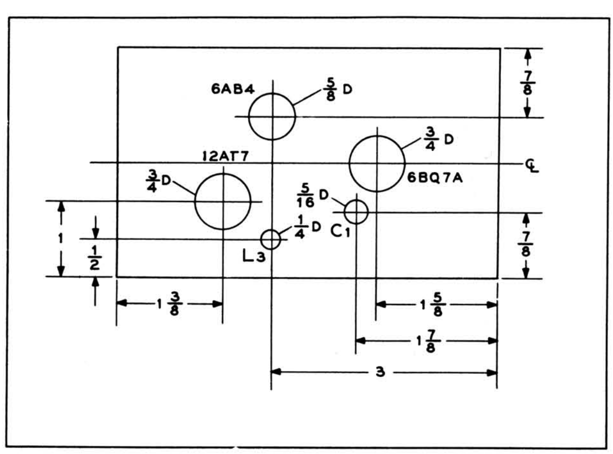


Fig. 5 Chassis drilling template for 144 mc beginner's converter.

The "Simple Sixer" Converter for 50 Mc.

A low noise crystal controlled converter is a necessity for the serious 50 mc enthusiast. The excellent noise figure of the cascode circuit proves to be a definite advantage during those early daylight or late evening hours when tropospheric-bending propagation is present, and ionospheric noise is at a minimum. A noise figure of approximately 4 db at 50 mc is produced by this converter. Heater and plate power can be obtained from any source capable of supplying 6.3 volts at 1.2 amperes, and 200 to 250 volts at 40 ma.

Converter Circuit

Shown in Figure 6 is the schematic of the "Simple Sixer" 50 mc converter. A 6BK7A cascode stage, employing a double-tuned input circuit is used as an r-f amplifier. The selective grid circuit is made from a single, tapped length of *B & W Miniductor* coil material, with the ends connected to the stator sections of a *Johnson type M* midget butterfly variable capacitor. A similar circuit is employed in the plate circuit of the cascode stage. These bandpass circuits are stagger-tuned to provide a flat-topped response curve about 4 mc wide. The excellent skirt selectivity of these double-tuned circuits help to prevent image interference in the 10 to 14 mc intermediate frequency range, and also reject VHF images which are troublesome in converters of this type.

The pentode half of a 6U8 acts as a mixer, with the triode section functioning as a harmonic crystal oscillator, using the fifth overtone of an 8 mc crystal to provide a 40 mc mixing signal. Crystal feedback may be regulated by the pi-network (L4-C4-C5), avoiding the necessity of using the more costly VHF crystals.

A second 6U8 serves as an i-f amplifier and a cathode follower output stage. The two i-f circuits (L5 and L6) are resonated by residual circuit

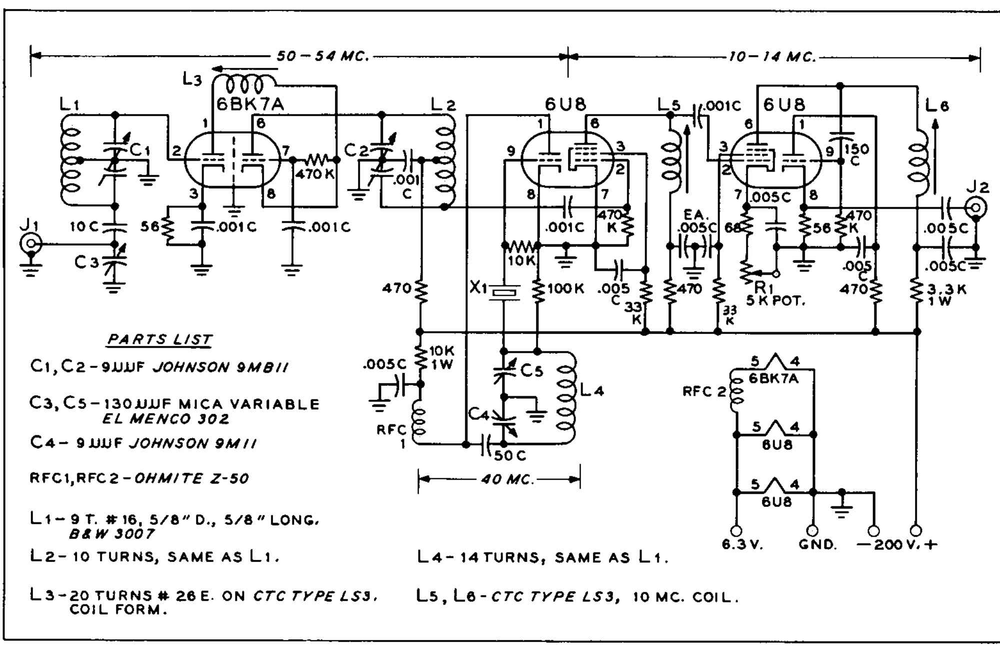


Fig. 6 Schematic, "Simple Sixer" converter for 50 mc.

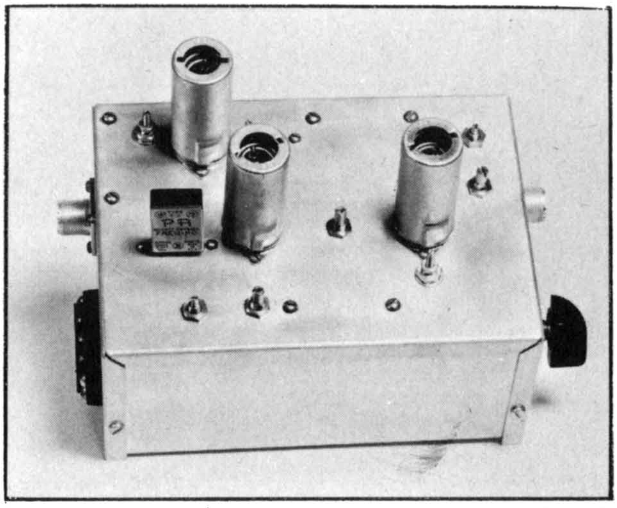


Fig. 7 Simple six meter crystal controlled converter that digs right down into the external noise level picked up by your antenna. 6BK7A r-f stage at right provides N/F of about 4 decibels. At left are 6U8 overtone oscillator and 6U8 i-f amplifier and cathode follower output stage.

capacities, and are stagger-tuned for improved band-pass performance in the 10 to 14 mc region. The i-f stage also provides sufficient gain so that this converter may be employed with receivers having no r-f stage. The gain control (R1) permits the converter output level to be adjusted for best performance without receiver overloading.

Converter Assembly

Top and bottom views of the "Simple Sixer" are shown in Figures 7 and 8. The converter is built upon one side of a 3″x5″x7″ aluminum box-chassis *(L.M.B. #145)*. The gain control (R1) and antenna input receptacle (J1) are located on one end plate of the box, and the i-f output receptacle

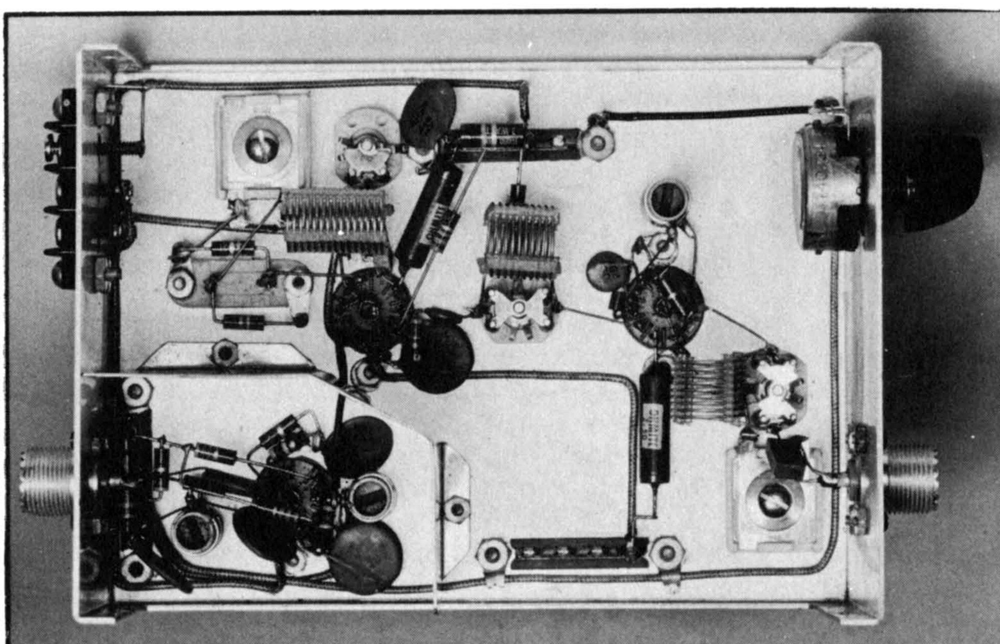


Fig. 8 Bottom view of the "Simple Sixer" 50 mc converter showing the placement of the shield around the i-f amplifier, and positioning of the r-f coils. 6BK7A cascode amplifier is at right. 6U8 mixer at left.

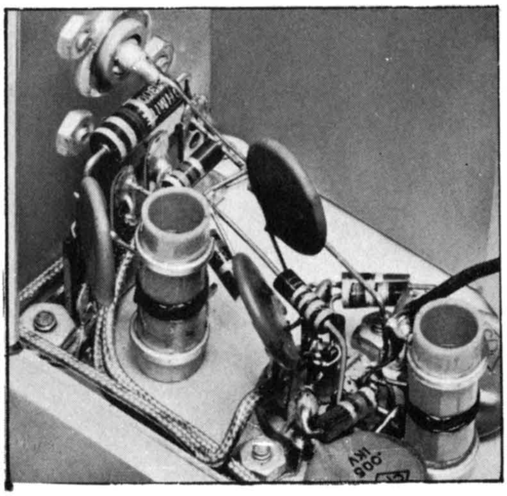


Fig. 9 Intermediate frequency amplifier compartment. At left is output coil, L6 and the coaxial receptacle. At right is the mixer plate coil L5, with plate lead of 6U8 tube passing through hole in the shield partition.

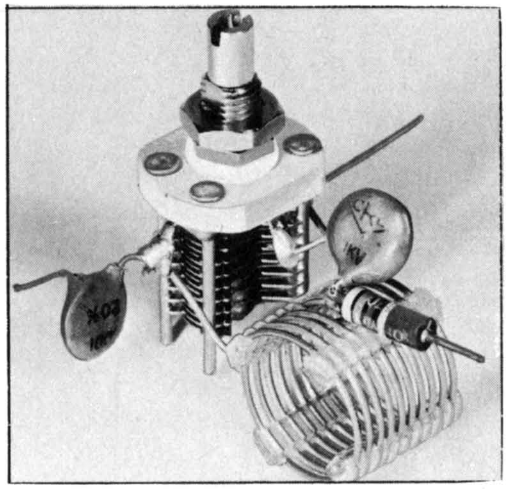


Fig. 10 Detail view of an r-f bandpass transformer, made from air inductor and a midget butterfly air tuning capacitor. Good skirt selectivity of the double tuned tank helps to prevent images.

(J2) and a 3-screw terminal strip occupy the other end plate of the box.

A small aluminum shield (visible in Figure 8) measuring 4¾"x3⅛" separates L5, L6, J2 and the i-f amplifier tube socket from the VHF circuits. A close-up of the i-f compartment is shown in Figure 9. The lead from plate pin 6 of the 6U8 mixer tube passes through a ¼" hole in the partition.

All resistors except the cathode bias and grid-to-ground units mount on the three 4-terminal *Cinch-Jones 2000-4* mounting strips placed at convenient locations. Bypass and coupling capacitors fasten directly on their associated parts, and to ground lugs placed under all the 4-40x¼" long machine screws holding the tube sockets, coaxial connectors, and terminal strips to the chassis. Heater, plate power, and gain control connecting leads run near the corners of the chassis.

The band-pass coupling transformers L1-C1 and L2-C2 are made from standard *Miniductor* coil material and midget butterfly capacitors, as shown in Figure 10. The coil is tapped by bending a coil turn each side of the tap enough to prevent it from becoming shorted when the connecting lead is soldered to the coil. All r-f and bypass connections should be made with shortest possible connections. Placement of under-chassis components may be seen in Figure 8.

After preliminary adjustments have been made, the low impedance output circuit of the converter is attached to the station receiver by a length of coaxial cable, and the coils L5 and L6 are peaked at 11 mc and 13 mc respectively, using a low level r-f signal fed into pin 2 of the 6U8 mixer socket.

When noise figure adjustments have been completed, a N/F in the region of about 5 db may be expected.

A "Very Low Noise" Converter for 144 Mc.

The cascode circuit provides an excellent noise figure with a minimum of parts. When used with high conductance VHF triodes, noise figures of 3 to 4 decibels at 144 mc may be obtained. One of the best triodes for use in the cascode circuit is the Western Electric 417A (5842), having a transconductance of about 25,000. Two of these tubes in the circuit of Figure 12 provide a N/F of approximately 3 db at 144 mc. The less expensive 6AM4 TV-type triodes may be substituted for the 417A's with a drop of N/F of about 1.5 db.

Converter Circuit

The complete schematic of the "Very Low Noise" 144 mc converter is shown in Figure 12. Two 417A tubes are employed in a broadly-tuned 144 mc cascode r-f amplifier. The input and output circuits of this stage are self-resonant at 145 mc, providing coverage of the lower 2 mc of the 2 meter band. Complete band coverage may be obtained, if desired, by stagger tuning these circuits, with only a slight drop in overall gain and noise figure. The cathode circuit of the first 417A r-f amplifier is series resonated to ground by C2 to enhance the noise figure. The cascode amplifier is neutralized by means of slug-tuned coil L2.

The second 417A tube is connected in grounded grid fashion, with the r-f signal injected into the cathode circuit. Filament chokes are used in this stage to prevent undesirable coupling between the cathode and filament circuits. The output of the cascode amplifier is inductively coupled to a triode connected 6AK5 mixer, which is capacitively coupled to a 6C4 inter-

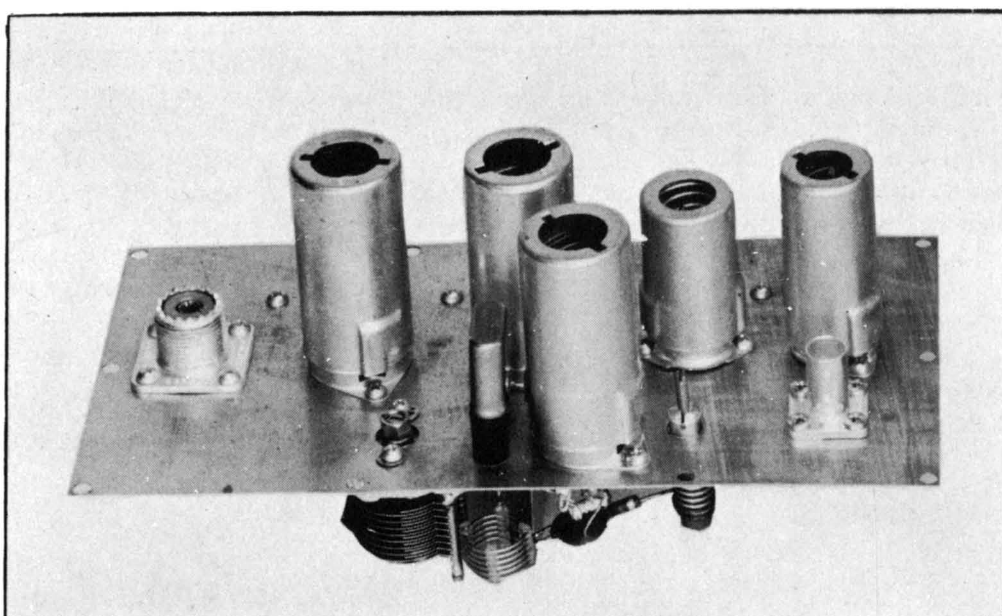


Fig. 11 The "Very Low Noise" 144 mc converter employs two 417A grounded grid triodes in a cascode r-f amplifier to give a N/F of approximately 3 db. The converter is constructed upon a brass plate mounted upon an inverted aluminum chassis which serves as a cabinet and shield. R-F stages are to the left, and the 12AT7 oscillator-multiplier is at front.

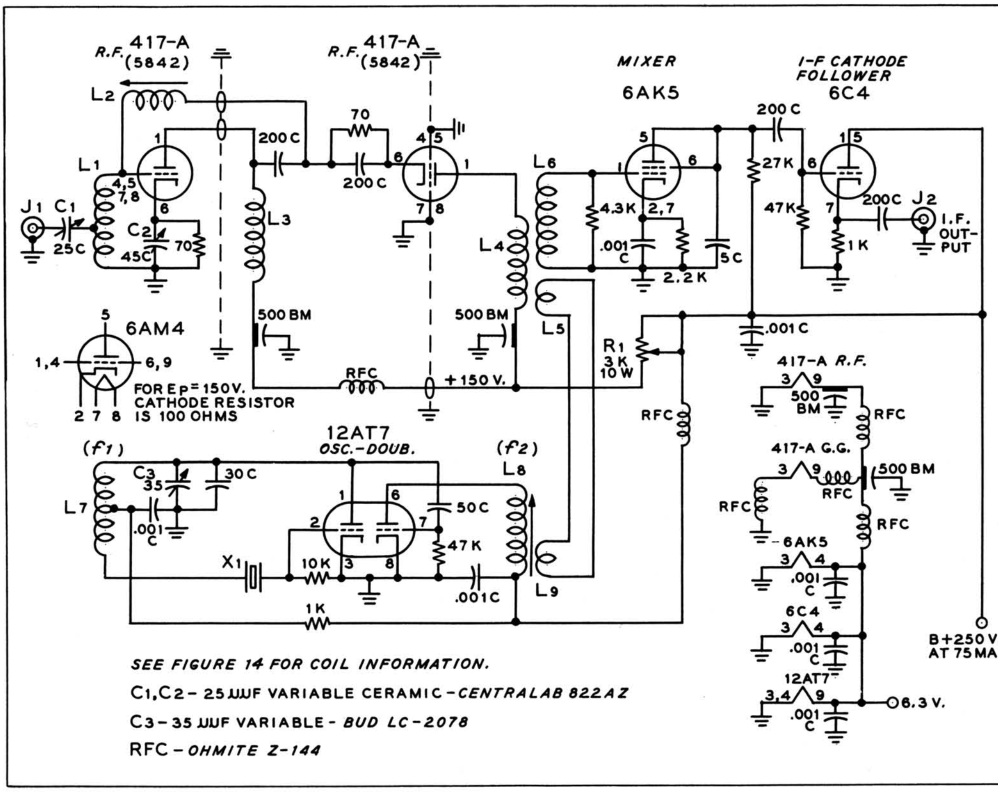


Fig. 12 Schematic, "Very Low Noise" converter for 144 mc.

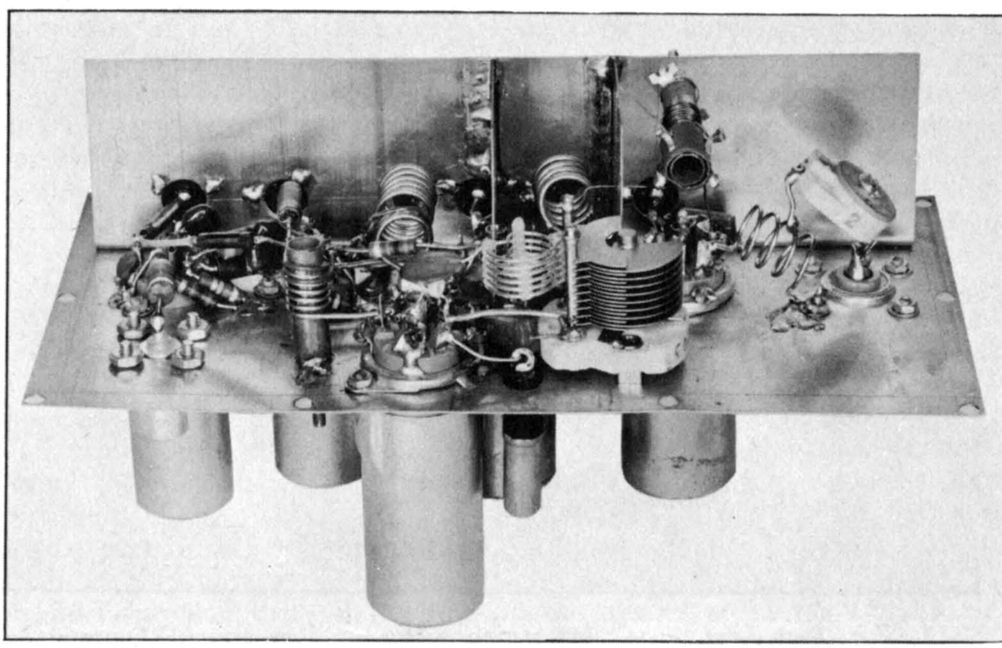


Fig. 13 Under-chassis view of 417A converter. Brass shields are placed between the r-f stages. The cascode neutralizing coil is mounted to the r-f compartment wall. Power leads pass through mica disc capacitors in shield to r-f filters.

COIL TABLE	I-F RANGE CHART			
	CRYSTAL (X 1)	f_1 (MC.)	f_2 (MC.)	I.F. (MC.)
L1 - 4 TURNS # 16 TINNED, 1/2" LONG, 3/8" DIAM. TAP APPROX. 1 1/2 TURNS FROM GROUND END.	13.0 MC.	65	130	14-18
L2 - 9 1/2 TURNS # 22 E., 1/2" LONG, 1/4" DIA. (SLUG-TUNED)	21.66 MC.	65	130	14-18
L3 - 5 TURNS # 16 TINNED, 1/2" LONG, 7/16" DIA.	43.33 MC.	43.33	130	14-18
L4 - 6 TURNS # 16 TINNED, 1/2" LONG, 7/16" DIA.	65.0 MC.	65	130	14-18
L5 - 1 TURN # 16 HOOKUP WIRE	9.46 MC.	47.33	142	2-6
L6 - 3 TURNS # 16 TINNED, 1/2" LONG, 7/16" DIA.	15.77 MC.	47.33	142	2-6
L7 - 5 1/2 TURNS # 14 TINNED, 3/8" LONG, 1/2" DIA. (*B&W 3003 MINIDUCTOR*) TAP 3 1/2 TURNS FROM PLATE END.	47.33	47.33	142	2-6
L8 - 5 1/2 TURNS, 1/2" LONG, 1/4" DIA. (SLUG-TUNED)				
L9 - 2 TURNS # 16 HOOKUP WIRE				

Fig. 14 Intermediate frequency chart for "Very Low Noise" converter.

mediate frequency low impedance cathode follower output amplifier stage.

Mixing voltage for the 6AK5 is derived from a link coupled 12AT7 oscillator-multiplier, employing an overtone crystal whose frequency is determined by the choice of intermediate frequency, as shown in the i-f range chart of Figure 14. The first section of the 12AT7 double triode acts as a third or fifth overtone oscillator, while the second section is a frequency doubler or tripler, whose plate circuit is link coupled to the grid circuit of the converter stage.

The "Very Low Noise" converter operates at a plate potential of 250 volts, drawing approximately 75 ma. Plate voltage for the cascode stages is dropped to 150 volts by means of the adjustable voltage control resistor, R1.

Converter Assembly

The physical arrangement of this converter is shown in Figures 11 and 13. The chassis of the unit is made of a thin piece of brass sheet, measuring 4" x 6". This sheet is mounted upon an inverted 2" aluminum chassis of the same dimensions as the brass sheet *(Bud AC-430)*. The aluminum chassis forms an enclosing shield for the under-chassis components. Placement of the major components beneath the chassis may be seen in Figure 13.

Grid pins 4, 5, 7 and 8 of the first 417A r-f tube socket are jumpered together by means of a short piece of copper strap. Grid coil L1 attaches to this strap, and is grounded at the opposite end to the frame of the coaxial

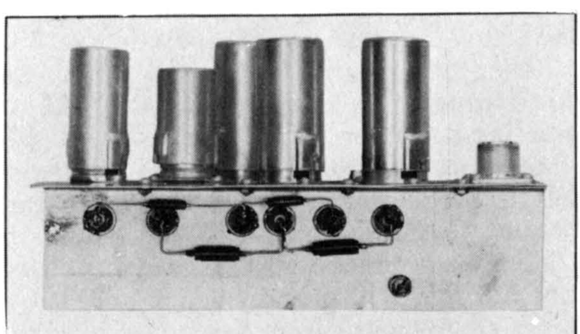


Fig. 15 Side view of chassis of converter. Note power lead decoupling chokes.

antenna receptacle, J1. Neutralizing coil L2 is mounted to the brass partition which runs the full length of the chassis.

Plate coil L3 of the first r-f stage, and the components of the second r-f stage are mounted in a compartment formed by two 2″ x 2″ brass plates soldered to the chassis, and to the full-length partition. The plate lead of the first 417A, and the lead from neutralizing coil L2 pass through two 1/4″ holes drilled in the shield partition. Grid pins 4, 5, 7 and 8 of the second 417A socket are grounded to the center stud of the socket, which in turn is grounded to the shield partition which passes over the center of the socket. A cut-out is made in the partition to clear the socket, permitting plate pin 1 and filament pin 3 to project into the left hand area, as shown in Figure 13.

The filament and plate power leads from the two 417A tube sockets pass through the full-length partition, utilizing *Centralab type ZB* rim mounting feedthrough mica capacitors mounted in holes in the brass partition. The r-f chokes incorporated in the power leads are mounted on the back of this partition, as shown in Figure 15.

The r-f plate coils are air-wound and are self supporting. They are mounted between the socket pins and the plate circuit feedthrough capacitors. Coil L6 is connected between pin 1 of the mixer socket and ground. Inductive coupling is employed between the frequency multiplier and the converter stage. Coupling between L5 and L6 may be varied for best gain and noise figure, as explained later.

If desired, 6AM4 tubes may be substituted for the 417A's, as indicated in Figure 12, with but a slight drop in noise figure.

Before power is applied to the converter, resistor R1 should be set to maximum value. When the oscillator, mixer and cathode follower stages are working properly, R1 should be adjusted for a potential of 150 volts measured at the plates of the 417A tubes. Each tube draws 25 milliamperes, and operates with 1.7 volts of cathode bias.

After N/F measurements have been made, the series cathode capacitor C2 should be adjusted to enhance the noise figure. It is wise to experimentally replace this unit with a 500 uufd disc mica capacitor to see if an improvement in N/F may be obtained.

A Low Noise 144-220 Mc. Grounded-grid Converter

This 220 mc converter uses the 6AJ4 and 6AM4 tube types developed especially for high frequency TV service. The N/F of the unit is about 5½ db, which compares favorably with more expensive and elaborate circuits. As shown in Figure 16, two stages of r-f amplification are used. These use 6AJ4's in grounded grid configuration. A 12AT7 is used as a conversion oscillator, the first section operating at 70 mc with an overtone crystal in a Butler-type circuit. The second half of the 12AT7 triples to 210 mc, and this frequency is coupled into the 6AM4 mixer by means of a 1 uufd coupling capacitor (C7). The resultant intermediate frequency varies from 10 to 15 mc, depending upon the frequency of the signal being received. The 200 to 225 mc range is thus covered by tuning the station receiver between 10 and 15 mc.

The two grounded grid stages are followed by a 6AM4 mixer stage. This tube has five grid pins, of which only two are used. The three unused grid pins were removed from the socket to minimize stray capacitance. A single

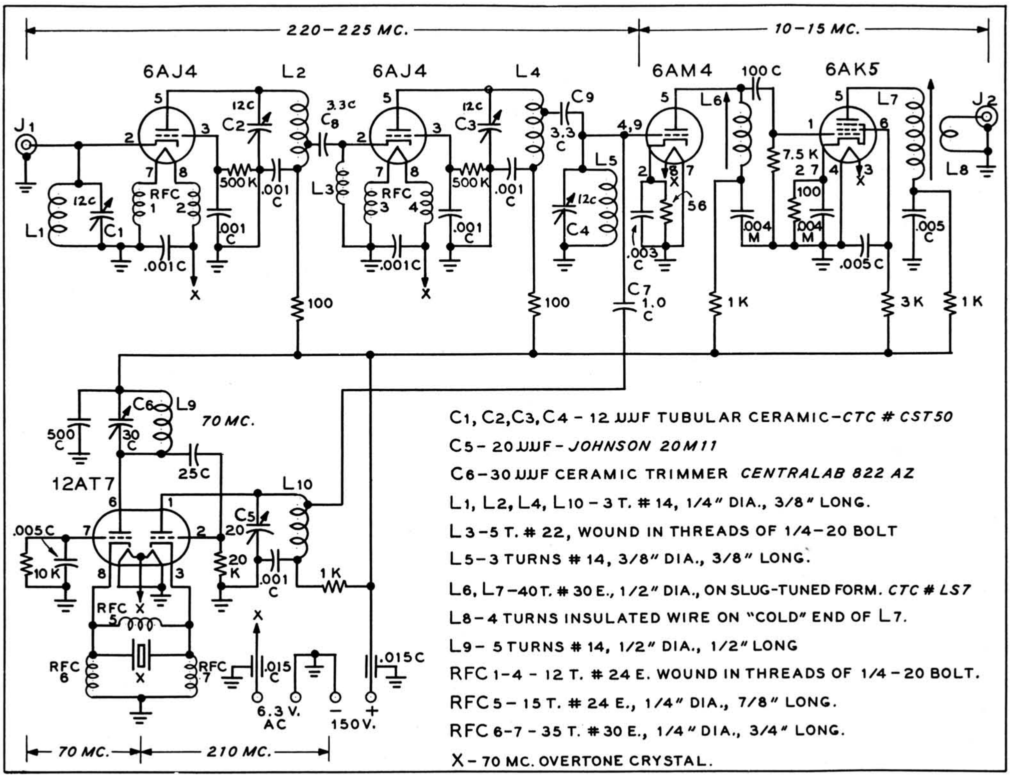


Fig. 16 Schematic, 144-220 mc grounded-grid converter.

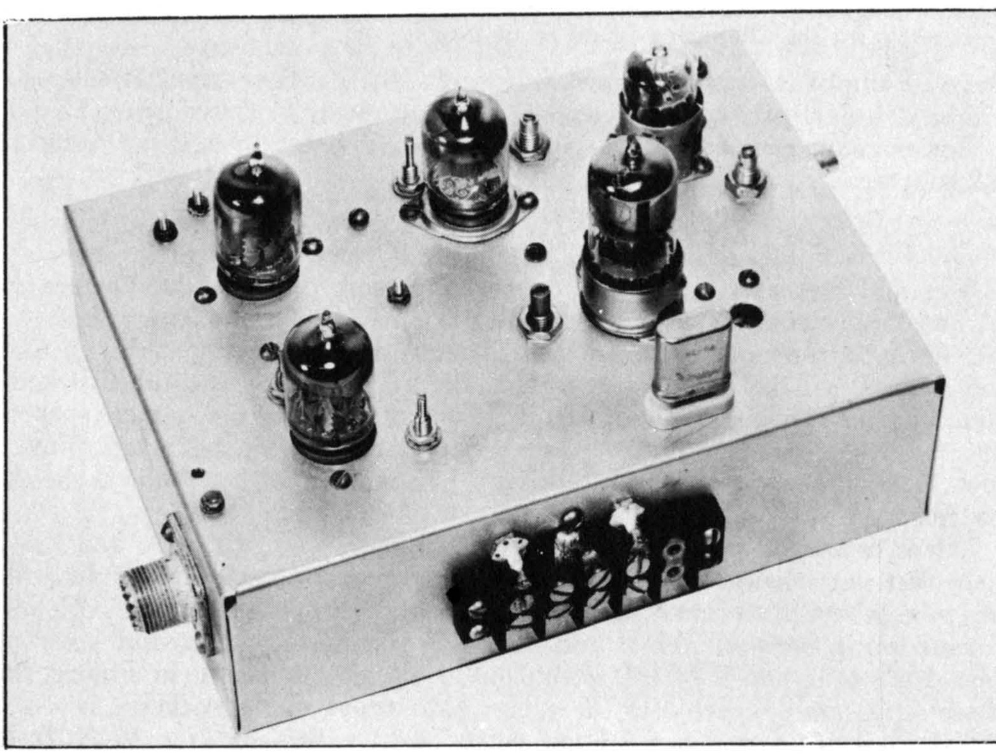


Fig. 17 This low-noise converter is designed for 144 mc or 220 mc use. Two 6AJ4 grounded grid r-f stages are at left, and overtone oscillator at right.

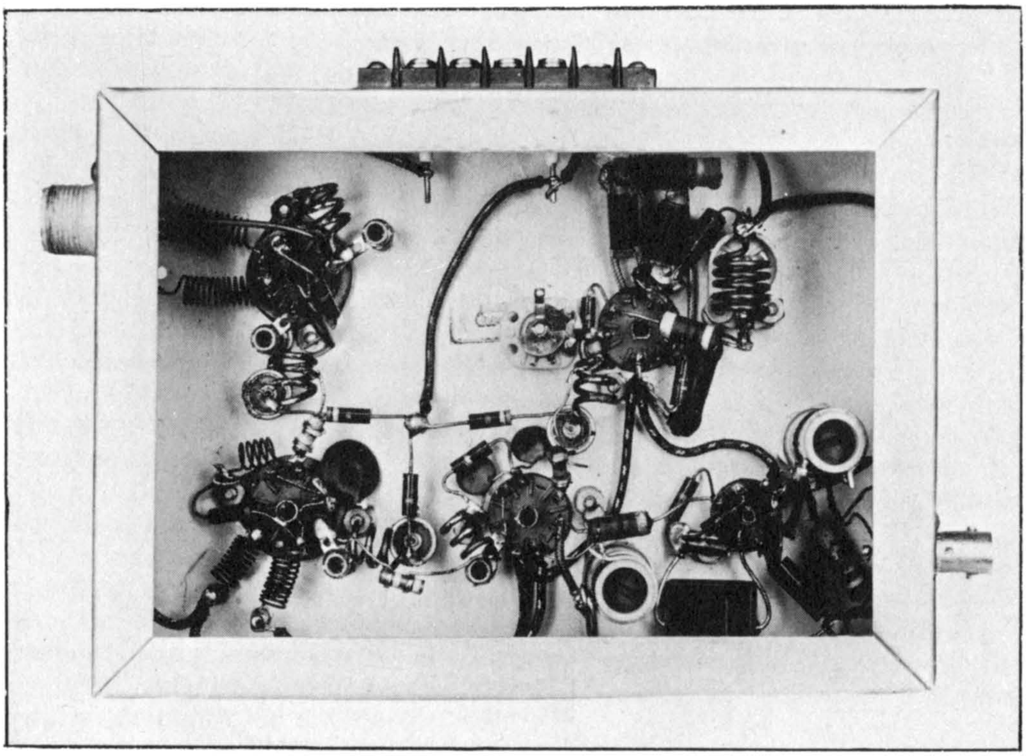


Fig. 18 Bottom view of 220 mc converter. R-F input at top left, second r-f amplifier at bottom left corner; mixer tube and slug-tuned plate coil at bottom center; and i-f amplifier and i-f output coil at bottom right. At top right is crystal socket, the 12AT7, and other components of the oscillator circuit. Terminal strip for all power leads is at top.

stage i-f amplifier uses slug tuned coils and a 6AK5. The output of this amplifier is inductively coupled to the station receiver by transformer L7-L8.

Power requirements are 150 volts at 50 milliamperes, and 6.3 volts at 1.2 amperes.

Converter Assembly

Top and bottom views of the converter assembly are shown in Figures 17 and 18. The converter is built upon a 5″x7″x2″ aluminum chassis *(Bud AC-402)*. Care should be taken to drill tube socket mounting holes so that the sockets will be properly oriented. Before mounting the tube sockets, some of the unused pins of the first r-f stage and the mixer sockets are removed to reduce stray capacity. In the 6AJ4 first r-f stage, pins 1 and 3 should be removed, and in the 6AM4 mixer stage pins 3 and 6 should be removed.

When mounting the tubular ceramic trimmers (C1, C2, C3, and C4), note that they should be oriented so their lugs can be soldered directly to the tube socket pins. Small strips of copper about 3/16″ wide are fashioned for soldering between pins 4 and 9 of the first 6AJ4 socket, and between pins 1, 3, 4, 6, and 9 of the second 6AJ4 socket. As shown in Figure 18, these strips are connected to the center base sleeve of the socket.

R-F chokes 1-4 and coil L3 are wound on the threads of a ¼-20 bolt. After the coils are completed, it is possible to carefully unscrew the bolt from the coils without disturbing the spacing. It is wise to leave the coil

leads longer than necessary until ready to wire the coils into the circuit.

The filament leads, r-f chokes, and high voltage leads should be wired first. Wiring of the main components should follow, with the exception of C7, C8, C9, and the lead from the input coaxial receptacle J1 to the cathode pin of the first 6AJ4. Care should be taken to keep all r-f leads as short as possible.

In wiring, note the placement of the capacitors on each side of pin 2 of the 6AM4 socket. Two capacitors are used here to reduce lead inductance. Connection of C7, C8, and C9 and the r-f input lead to their associated coils is somewhat critical for optimum noise figure, and should be adjusted as described below.

The adjustment procedure for obtaining the best noise figure involves moving the connections of the r-f input lead, C7, C8, and C9. Of course, before these adjustments can be made, these components must be temporarily soldered in place to check voltages and to make sure the unit is functioning properly. It is suggested that as a starting point, the r-f input lead from J1 be soldered directly to the cathode pin of the socket of the first 6AJ4 r-f amplifier. The connection of C8 to L2 can be made at the center of L2. Then C9 can be soldered from a point about one-half turn down from the plate end of L4 directly to pin 9 of the 6AM4 tube socket. C7 can be connected between grid pin 4 of the 6AM4 socket, and a tap about one-half turn down from the plate end of L10.

With these connections made, the first step is to use a 220 mc test signal and align all tuned circuits for maximum gain. Once that is accomplished the objective of subsequent adjustments is to improve the noise figure. A typical test set-up for noise figure measurements is discussed in chapter 12 of this Handbook.

In the model shown J1 was tapped at the cathode pin of the tube, and C7 was tapped ½-turn down from the plate end of L10. C8 was tapped at the center of L2. Adjustment of C9 proved to be noncritical.

144 Mc. Operation

With modifications to the tuned circuits, this converter may be made to operate on 144 mc. A measured noise figure of better than 5 db may be obtained at this frequency. Necessary coil and circuit modifications are tabulated in Figure 19.

Alignment of the converter on 144 mc follows the same procedure as

Fig. 19 144 mc coils for grounded-grid converter of Figure 16.

DATA FOR 144 MC. OPERATION

L1—4 T. #16, 1/2" DIA., 5/8" LONG. ANTENNA TAP 1.5 TURNS FROM GROUND END. CATHODE TAP 2.8 TURNS FROM GROUND END.

L2, L4—SAME AS L1. CATHODE TAP 1.3 TURNS FROM GROUND END.

L3—13 T. #24, WOUND IN THREADS OF 1/4-BOLT.

L5—4 TURNS #16, 3/8" DIA., 5/8" LONG.

L6, L7—SAME AS L6, L7, FIGURE 16.

L8—SAME AS L8, FIGURE 16.

L9—FOR 44.667 MC. CRYSTAL; 6 T. #20, 1/2" D., 5/8" L.

L10—3 T. #16, 1/2" DIA., 5/8" LONG. C7 TAPPED 1.5 TURNS FROM "COLD" END.

X—44.667 MC. OVERTONE CRYSTAL.

that discussed for 220 mc. The 6AM4 mixer stage is tested first, with L7 peaked at 10.5 mc, and L6 peaked at 12.5 mc.

The 12AT7 oscillator tube and a suitable crystal are next inserted, and a 0-25 ma meter temporarily connected in the plate voltage lead to L9. A sharp dip in plate current is noted when C6 is tuned to the crystal frequency. Oscillator operation should be checked with a receiver tuned to 44.667 mc. If a series of oscillations are noted, RFC-5 should be reduced in size a turn at a time until a single crystal controlled oscillation occurs.

The r-f amplifier section may be adjusted either for a nearly flat 144-148 mc bandpass, or the 144-146 mc portion of the band may be favored. Full bandwidth is obtained by adjusting C4-L5 to 145 mc, C3-L4 to 146.5 mc, C2-L2 to 146 mc, and C1-L1 to 144.5 mc respectively. The external signal source, and not the converter noise should be peaked at each specified frequency. Finally, the taps on coils L1, L2, L4, and L10 are adjusted for the best noise figure as in the manner described for 220 mc operation.

A Grounded-grid Converter for 420 Mc.

One of the simplest and best amplifiers for the 400 to 500 mc region is the grounded grid amplifier. When two stages of 6AM4 g-g triodes are employed, a noise figure in the region of 6 to 7 db may be obtained. If 417A tubes are substituted for the 6AM4's, an improvement of about 1½ db can be obtained over the original noise figure.

Converter Circuit

The converter described herewith employs two 6AM4 tubes as grounded

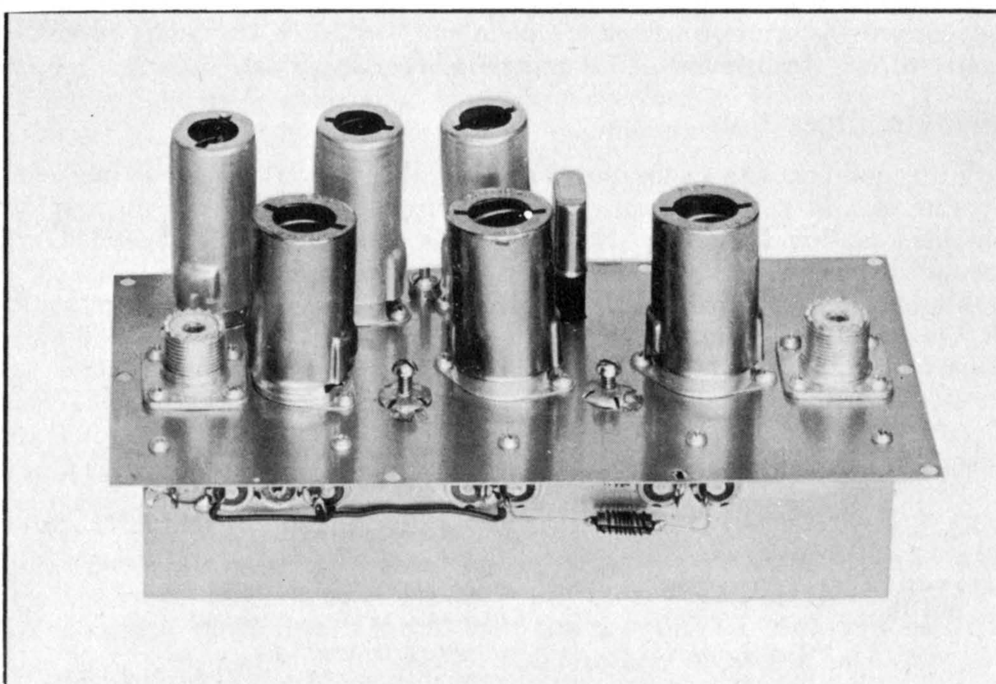
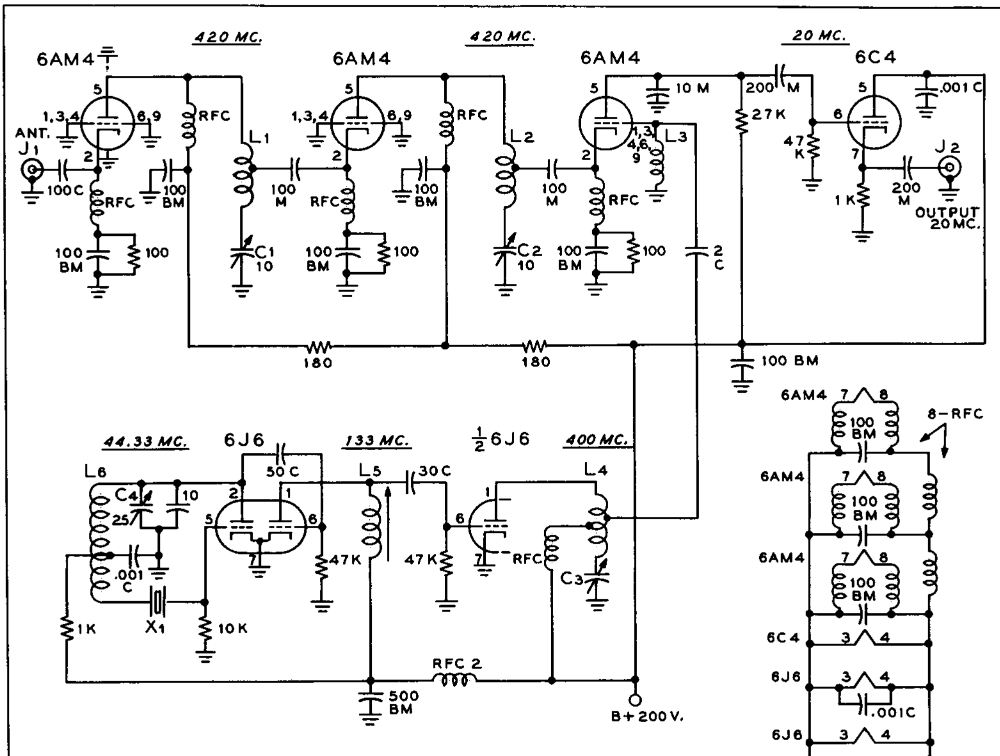


Fig. 20 Two stages of 6AM4 grounded grid triodes provide a N/F of 6 to 7 db at 420 mc in this simple converter. R-F tubes are in foreground, and harmonic oscillator, doubler and cathode follower tubes are at rear of chassis.



C1, C2, C3 - *ERIE 832-10*

RFC- *OHMITE Z-420*

RFC2 - *OHMITE Z-235*

L1, L2 - 1 1/2 TURNS # 12, 7/16" DIA., 3/8" LEADS. TAP 1/4 TURN FROM PLATE END.

L3 - 1" OF # 12 WIRE, CONNECTED FROM SOCKET STUD TO SHIELD.

L4 - 1 1/2 T. # 12, 7/16" D., 3/4" LEADS. INJECTION TAP 1/4" FROM PLATE OF 6J6. B+TAP AT CENTER.

L5 - 4 TURNS # 18, 1/4" DIA., 1/2" LONG ON IRON SLUG FORM.

L6 - 6 TURNS # 16, 5/8" DIA., 3/8" LONG. TAP ONE TURN FROM GRID END.

X1 - 44.33 MC. CRYSTAL.

Fig. 21 Schematic, 420 mc grounded-grid converter.

grid amplifiers, and a third 6AM4 as a mixer. A 6J6 double triode is used as an overtone oscillator and harmonic amplifier. One-half of a second 6J6 is used as a frequency multiplier, delivering 400 mc mixing voltage obtained from the 44.33 mc overtone oscillator. A 6C4 triode is used as an intermediate frequency cathode follower, coupling the 20 to 40 mc output of the converter to the station receiver. The i-f range of the converter may be modified as desired by changing the frequency of the mixing voltage.

The low impedance cathode input circuit of the first g-g amplifier is designed to match a resonant antenna coupled to the converter by means of a 52 or 72 ohm coaxial transmission line. The plate circuit of the first r-f stage is a balanced, series tuned assembly, with C1 approximately equal to the output capacity of the 6AM4 tube. A similar coupling circuit is used between the second g-g r-f stage and the 6AM4 mixer.

Grid injection is employed in the mixer stage. The various grid pins of the 6AM4 socket are strapped to the center pin of the socket, and grounded to the chassis by means of a 1¼″ length of #12 copper wire (L3). The inductance of this connection is sufficient to develop a satisfactory level of

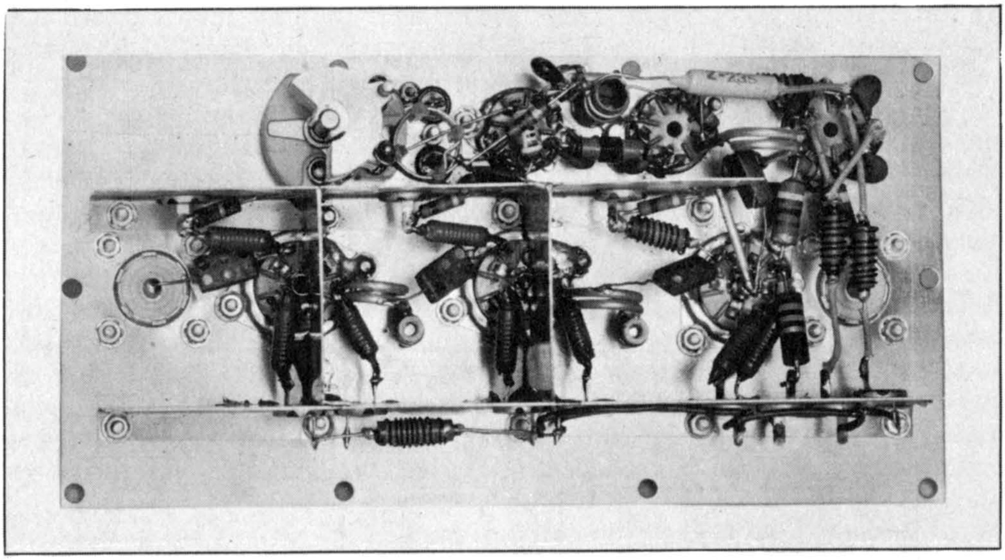


Fig. 22 Under-chassis view of 420 mc converter. First g-g amplifier stage is at left, with brass shield across center of tube socket. 6AM4 mixer grid inductor (L3) is visible in right-hand compartment. Oscillator-multiplier string is visible along top edge of chassis.

mixing voltage in the mixer circuit.

The plate circuit of the 6J6 400 mc tripler stage employs a series tuned circuit, similar to that in the plate circuit of the two r-f amplifier stages. Mixer injection level is controlled by varying a tap on coil L4.

Converter Assembly

The complete converter is built upon a brass plate measuring 4″x6″ which fits over the open side of an aluminum chassis measuring 4″x6″x2″ *(Bud AC-430)*. The under chassis area of the converter is broken up into five compartments, as shown in Figure 22. The two main partitions are cut from thin brass stock, and are bolted to the chassis by means of 4-40 machine screws. Two smaller brass shields pass across the 6AM4 tube sockets, and are grounded to the center stud of each socket. The bottom edge of each of these partitions is notched to pass over the socket, and all grid terminals of each r-f stage are grounded to this partition, and to the center stud of the socket. These two shield plates are soldered to the main shield partitions.

100 uufd *Centralab type ZA* mica feedthrough capacitors are mounted in the longer of the side partitions to pass the filament and plate power leads from the r-f compartments to a separate shielded area.

The first 6AM4 socket is oriented with pins 1 and 9 facing the coaxial antenna receptacle, J1. The shield partition passes across the socket between pins 3 and 4, and 6 and 7. The cathode, and filament r-f chokes are located in the grid compartment, and the plate r-f choke is located in the grid compartment of the second r-f stage.

The second 6AM4 socket is positioned in the same manner as the first, while the 6AM4 mixer socket is oriented so that pin 1 faces the preceding r-f stage socket.

Placement of parts in the crystal oscillator-multiplier string is not critical,

except that coil L4 of the last tripler stage should be oriented as shown in the bottom view photograph of Figure 22.

The coils for the 400 mc region are wound out of #12 silvered copper wire. Coils L1, L2 and L4 take the form of hairpins, having a 1½ turn loop at the top end. Coil L3 is merely a short length of this wire, reaching from the center stud of the mixer socket to the side shield partition.

Wiring of the r-f stages should be done first. The r-f chokes mount between socket pins of the tubes and the 100 uufd bypass capacitors are affixed to the shield partitions. Coils L1 and L2 are placed in position after all other wiring has been done. All small components of the oscillator-multiplier string mount between tube socket pins and other components mounted on the chassis. Filament and plate decoupling r-f chokes are mounted between the pins of the 6C4 socket and feedthough capacitors in the adjacent shield plate.

In the model shown, best N/F was obtained with the interstage coupling capacitor tapped ¼-turn up from C1. The capacitor tap on L2 was found to be optimum about ¼-turn from the plate end of the coil. Mixer injection may be adjusted by varying the length of L3.

The "Gold Plated Special" Preamplifier

Without a doubt the 416B high transconductance planar triode is by far the best tube to use in the VHF region at the present state of the art. With a transconductance of approximately 50,000 the 416B may be considered to approach the "absolute" noise figure in the VHF bands. The noise figure of the preamplifier shown herewith is so low as to be difficult to measure, being well under 3 db on 144 mc, 220 mc and 420 mc. Since the measurement of noise figure is difficult, and the results are usually open to question if not performed under laboratory conditions, it is sufficient to state that the noise figure of this r-f amplifier is measurably better by several decibels

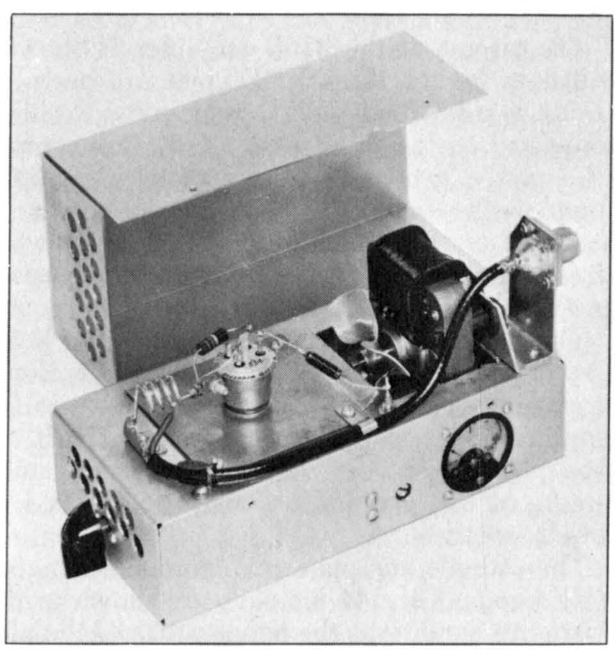
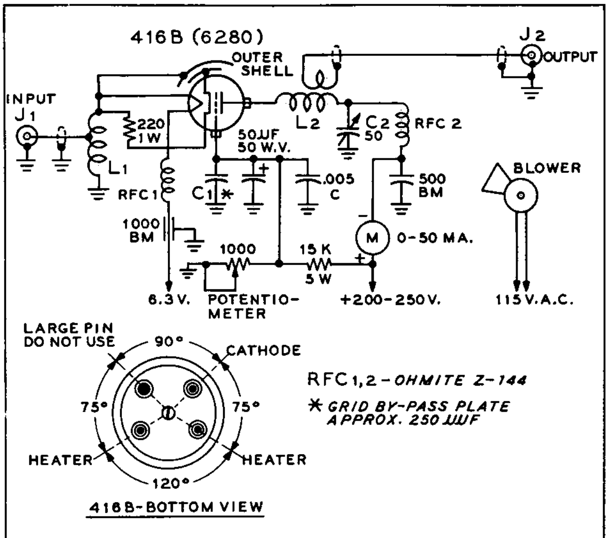


Fig. 23 "Ultimate" tube for VHF is the 416B, shown here in all-band preamplifier. Noise figure of better than 3 db is obtained up to 420 mc.



	144 MC.	220 MC.	420 MC.
L1	4 T. 3/8" I.D. SPACED 2 DIA. TAP 2 1/2 T. UP	3 T. 3/8" I.D. SPACED 2 DIA. TAP 2 T. UP	FLAT STRAP 1/2" WIDE 2" LONG TO CHASSIS. TAP ANT. NEAR CATH.
L2	4 T. 5/8" I.D. SPACED 1 1/2 DIA.	4 T. 3/8" I.D. SPACED 2 DIA.	FLAT STRAP 1/4" WIDE FROM STANDOFF TO VARIABLE CAPACITOR.

ALL COILS WOUND WITH # 14 BARE TINNED COPPER.
INPUT COIL SHOULD BE SQUEEZED AND TAP ADJUSTED FOR BEST NOISE FIGURE. OUTPUT LINK IS A 1 TURN LOOP, AND SHOULD BE ADJUSTED FOR OPTIMUM COUPLING AND CEMENTED IN PLACE.
ADJUST POTENTIOMETER FOR PLATE CURRENT OF ABOUT 25 MA.
CONNECTIONS TO CATHODE AND HEATER PINS ARE MADE WITH CONTACT CLIPS REMOVED FROM CONVENTIONAL MINIATURE SOCKET. GRID CONTACT RING IS 3/4" DIA. AND IS THREADED 40 TURNS PER INCH.

Fig. 24 Schematic, "Ultimate" preamplifier using high transconductance Western Electric 416-B planar triode. Noise Figure of better than 3 db is achieved on all amateur bands up to 420 mc. Tube must be forced-air cooled.

than any other receiver or amplifier tested to date. If the ultimate in receiver sensitivity is desired, this amplifier meets all requirements.

Preamplifier Circuit

The circuit of the 416B amplifier is shown in Figure 24, together with coil data for the three VHF amateur bands. A hot cathode, grounded-grid circuit is used. The 416B is well suited to this application, as the cathode is connected to the outer shell of the tube, providing separate terminals for r-f and d-c returns. Under normal conditions, the 416B operates under slight positive grid bias, with a plate current of 25 milliamperes at a plate potential of 200 to 250 volts. A bias network has been included in this circuit to allow the plate current to be adjusted to the correct value, as read on the 0-50 d.c. plate milliammeter incorporated in the preamplifier.

The hot cathode input impedance of the 416B is of the order of 250 ohms. The low impedance coaxial input is therefore tapped on the cathode coil at a point providing the best value of noise figure. The plate circuit of the amplifier stage employs a split tank circuit, with a low impedance pickup loop placed near the center of the tank coil. The low impedance output circuit of the preamplifier may be coupled into the input circuit of the station receiver.

The cathode and plate coils must be changed for operation on the various VHF bands. The 144 mc coils are shown in the photographs. Under proper operating conditions the temperature of the plate and grid seals of the 416B

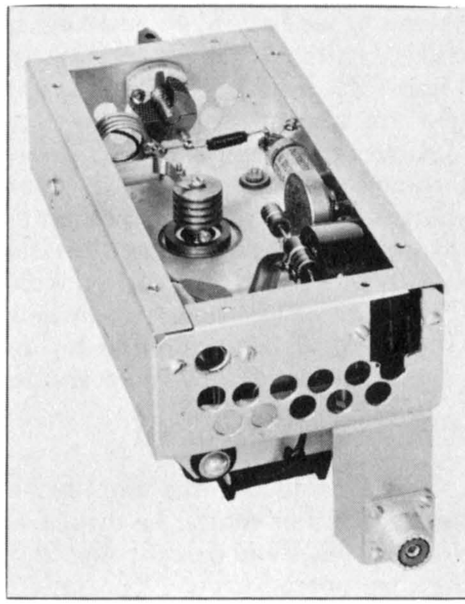


Fig. 25 Under-chassis view of "ultimate" preamplifier. Plate cap of 416B tube is visible at center, and the series plate tuning coil and capacitor are at rear of the chassis. Coaxial input receptacle is on bracket in foreground.

tube should not exceed 100 degrees Centigrade. Seal temperature is held below this value by a stream of air directed at the tube from a small 115 volt blower motor. The filament of the tube should never be turned on without first starting the blower, and the blower should be left running for five minutes after filament and plate voltage have been removed from the 416B tube.

Amplifier Assembly

The amplifier and associated blower are mounted on an aluminum chassis measuring 4″ x 8″ x 2″. A cutout is made at one end of the chassis to accommodate the blower motor and fan blade. The motor is mounted with the shaft in a horizontal position on a level with the chassis top. The edge of the chassis in front of the fan is bent up at a slight angle to deflect a portion of the air stream downwards towards the plate cap of the 416B planar triode tube. The motor is shock mounted to reduce vibration.

A 1¼″ hole is cut in the chassis for the 416B, which is mounted vertically, with the plate connection projecting into the under-chassis area. The tube is held in position by the grid mounting capacitor, C1. This low inductance capacitor is fabricated from a brass plate measuring 3″ on a side. It is separated from the chassis by a sheet of waxed paper. This paper provides sufficient voltage insulation, since the potential difference between the brass plate and the chassis is only 8 volts or so. The grid capacitor plate is held to the chassis by four 6-32 machine screws passing through ¼″ holes in the chassis. These screws are insulated from the chassis by fibre shims and washers. The capacity of this grid capacitor is approximately 250 uufd.

A hole is drilled in the center of the brass plate large enough to barely pass the ¾-40 thread of the grid ring of the 416B. The tube is held in position by means of a ¾″ nut placed on the opposite side of the grid bypass plate. Connections to the cathode and heater pins of the 416B are made with contact clips removed from a conventional miniature socket. The r-f connection to the cathode shell is made by a ¼″ wide copper strap

passed around the shell and held in place with a 6-32 nut and bolt. Cathode coil L1 attaches between this strap and a small ground clip, shown in Figure 23. The 220 ohm, 1-watt cathode resistor and one filament contact clip are also attached to this strap.

A transmitting-type heat dissipation connector is used for the plate terminal of the tube and is connected to the tuned circuit L2-C2 by a short length of ¼″ copper strap. The filament lead to the tube passes through a .001 ufd ceramic feedthru capacitor mounted on the chassis deck and then through filament choke, RFC-1.

A series of ⅜″ holes are drilled at each end of the chassis to permit free passage of air. If desired, a top cover may be placed over the amplifier to protect the 416B tube from accidental damage.

Amplifier Adjustment

Upon completion the amplifier should be inspected for wiring errors and tested. The fan should be turned on, and filament and plate voltage applied to the 416B. Plate current should be adjusted to 25 milliamperes by varying the bias potentiometer. The input and output circuits should be resonated to the operating frequency. Finally, the input coil should be squeezed and the antenna tap adjusted for optimum noise figure. The output link should be adjusted for optimum coupling to the receiver as a last step. Care should be taken to make sure that the blower motor is always running before filament power is applied to the tube. It should also be left running for five minutes after the filament of the 416B is turned off.

CHAPTER X

VHF Transmitter Design

Very High Frequency transmitters have progressed from the simple modulated oscillator of the "thirties" to the modern multistage transmitter of today. This change was accelerated by the discovery that stable transmitters and low noise, selective receivers held the secret to beyond-horizon communication on the VHF bands. The old fashioned modulated oscillator and the superregenerative "rush box" had neither the stability nor the selectivity for use with the narrow band communication circuits typical of the present state of the art.

The Modern VHF Transmitter

The modern VHF transmitter is basically similar to transmitters operating in the high frequency portion of the radio spectrum. The oscillator stage employs a crystal or stable v.f.o. for frequency control. The control frequency is raised to the operating frequency by a series of multiplier stages which drive the class-C amplifier stage of the transmitter. Suitable attention must be given to the suppression of unwanted harmonics and subharmonics generated by the multiplier string, otherwise except for the physically small tuned circuits and components the VHF transmitter resembles its low frequency counterpart.

Frequency Multiplication

A considerable latitude exists in the choice of the control frequency of the VHF transmitter. The oscillator may operate on the fundamental output frequency of the transmitter, or it may operate on some sub-harmonic of the output frequency.

Figure 1 shows fundamental frequency operation at 144 mc for second to twentieth harmonic operation. The third column lists the order of multiplication necessary to achieve this harmonic mode. Harmonics employing doublers (x2), triplers (x3), and quadruplers (x4) are commonly used for VHF work. Higher order frequency multiplication than this within a

CRYSTAL HARMONICS FOR 144 MC. OPERATION		
FUNDAMENTAL FREQUENCY, MC.	HARMONIC	FREQUENCY MULTIPLICATION
72.0	2	X 2
48.0	3	X 3
36.0	4	2 X 2, X 4
28.8	5	—
24.0	6	2 X 3
20.6	7	—
18.0	8	2 X 2 X 2, 2 X 4
16.0	9	3 X 3
14.4	10	—
13.1	11	—
12.0	12	3 X 2 X 2, 3 X 4
11.1	13	—
10.3	14	—
9.6	15	—
9.0	16	2 X 2 X 2 X 2, 4 X 4
8.48	17	—
8.0	18	2 X 3 X 3
7.57	19	—
7.20	20	—

Fig. 1 The first twenty crystal harmonics for 144 mc operation are listed in this chart. The prime harmonics (fifth, seventh, etc.) are unuseable, as are those harmonics requiring high values of frequency multiplication, such as the tenth, thirteenth, and fourteenth.

single stage results in such low plate efficiency that very little output is available from the circuit. Operation on the fifth, tenth, or fifteenth harmonic modes is therefore impractical. It should also be noted that certain harmonics are prime numbers, indivisible by any number other than themselves and unity. The fifth and seventh harmonics are prime numbers, and are therefore not recommended. In general, the sixth, eighth and twelfth harmonics are the most useful for VHF work, as they may be reached by easy multiplication steps.

The third and fifth crystal overtones may be utilized in harmonic oscillators, followed by doubler or tripler stages. In most cases, 8 mc, 12 mc, or 24 mc is chosen for the control frequency for 144 mc, since inexpensive crystals are available in these regions.

Frequency Multipliers

The frequency multiplier chain should be designed to deliver a maximum of desired harmonic voltage, with a minimum of unwanted harmonics. These incompatible ideals can only be combined by employing a sufficiency of high-Q, inductively coupled tuned circuits between stages. Simple capacitive coupling between multiplier stages permits spurious frequencies to pass easily through the stages with little or no attenuation. A compromise may be reached in amateur service between circuit complexity and maximum harmonic suppression, unless the VHF transmitter happens to be operating in an area of weak-signal TV reception. The use of an auxiliary antenna tuner will often suffice if it is desired to eliminate spurious VHF radiations from an existing transmitter having insufficient adjacent channel suppression in its tuned circuits.

Frequency Triplers

The use of push-pull frequency triplers is very popular in the VHF region, particularly for 144 mc and 420 mc operation. A push-pull amplifier designed for 50 mc operation will act as an efficient frequency tripler from 48 mc to 144 mc, requiring only a change of plate circuit constants. Like-

wise, a push-pull 144 mc stage can act as a frequency tripler from 140 mc to 420 mc. The push-pull tripler stage should not be used as the final stage of a transmitter, as a considerable amount of fundamental frequency energy appears in the plate circuit. When the tripler is employed to drive a succeeding amplifier, the tuned circuits of the amplifier will usually offer sufficient selectivity to prevent radiation of the sub-harmonic passed by the tripler.

TVI and Frequency Multiplication

It must be remembered that when low frequency oscillators are used to drive a chain of frequency multipliers many other harmonics than the desired one are generated. A doubler stage will produce third and fifth order harmonics, as well as a small amount of fundamental frequency energy. These undesired frequencies appear because a simple tuned circuit, such as employed in a frequency multiplier does not have sufficient selectivity to completely reject all the undesired frequencies produced by the multiplier.

In particular the 6.25 mc energy passed through the exciter can beat with the 50 mc signal and produce a spurious signal on 56.25 mc, right in the middle of television channel two. Second harmonic energy (12.5 mc), in like manner, can beat with the 50 mc signal, producing a spurious 62.5 mc signal, falling in television channel three.

In the case of a 144 mc transmitter, emission of a 72 mc sub-harmonic can interfere with the sound carrier of television channel four, and various beats and mixtures of the lower harmonic frequencies can fall within other TV channels.

The solution to this problem, then, is to restrict the radiation of unwanted harmonics, and to generate the carrier frequency at as high a frequency as practical, eliminating as many spurious sub-harmonic frequencies as is possible. In areas where TV channel two is in operation it is wise to place the control oscillator of a 50 mc transmitter on 25 mc, or possibly directly on 50 mc, eliminating the sub-harmonics that would fall within the limits of the TV channel. In similar fashion, 48 mc and 72 mc overtone crystals are becoming popular for 144 mc operation in areas of television reception.

VHF Oscillator Circuits

Several simple fundamental frequency crystal oscillator circuits are shown in Figure 2. The grid-plate triode oscillator has the advantage of simplicity, although the power output available without damage to the crystal is low. The pentode tuned plate oscillator requires more components, but is capable of greater power output with less strain upon the crystal when operated at an equivalent plate voltage. Crystal current may be measured by placing a 60 milliampere (pink bead) pilot lamp in series with the crystal, as shown. The lamp will light when the r-f crystal current approaches the maximum safe limit of 60 milliamperes, and will burn out before excess r-f current can fracture the crystal.

A more economical method of generating a stable signal in the VHF range is to employ a harmonic oscillator which will deliver output at some harmonic of the fundamental crystal frequency.

The *Tri-tet* harmonic oscillator is shown in Figure 2B. When used with

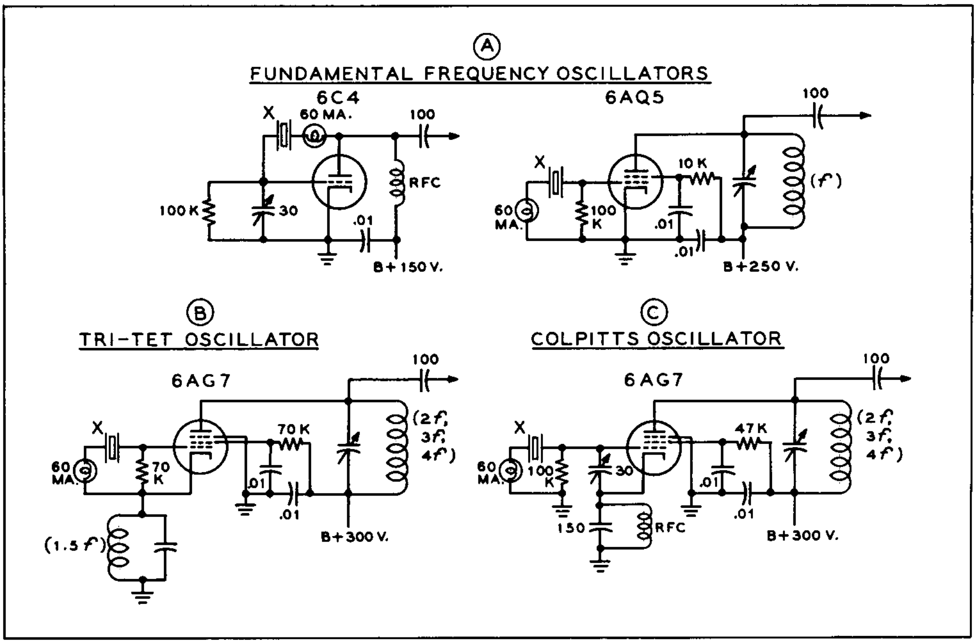


Fig. 2 Four basic crystal oscillator circuits suitable for VHF work are shown above. Tri-tet circuit delivers high output with low crystal current.

a high transconductance tube such as a 6AG7 or a 6CL6, high output is obtained up to the crystal fourth harmonic with a minimum value of crystal current. Crystals in the 7 mc - 9 mc region may be used to obtain output up to approximately 36 mc, and overtone crystals may be used to obtain output up to and above the 50 mc band. The resonant cathode circuit is fixed tuned to 1.5 times the crystal frequency for maximum harmonic output and minimum crystal current.

The *Colpitts* harmonic oscillator circuit is shown in Figure 2C. A capacitor voltage divider circuit between grid and cathode furnishes feedback the magnitude of which is controlled by the setting of the variable grid capacitor. R-f output and crystal current compare with those values obtained with the tri-tet circuit.

Overtone Oscillators

The new overtone crystals capable of operation on the third, fifth, or seventh modes have made possible the use of double triode regenerative oscillators, such as shown in Figure 3. To sustain oscillation it is necessary to supply a feedback path between the output circuit and the crystal circuit, otherwise the configuration of the overtone oscillator is similar to that of a simple fundamental frequency oscillator. The feedback path usually consists of a grid coil coupled to the plate tank circuit. This coil may be a part of the plate coil (as illustrated in Figure 3A), or it might take the form of a separate coil (Figure 3B). Capacitive feedback may also be employed, as shown in Figure 3C. Only sufficient feedback to maintain circuit oscillation should be used, as excessive feedback will lead to instability and possible crystal fracture.

The degree of feedback in the circuits of Figures 3A and 3B is determined by the number of turns on the oscillator coil between r-f ground potential

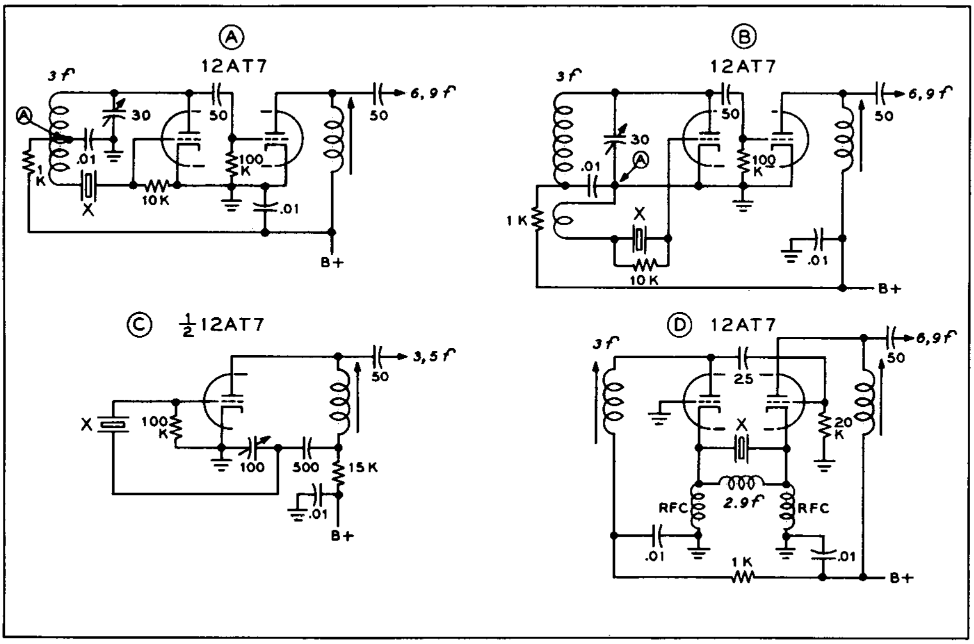


Fig. 3 Four VHF overtone crystal oscillator circuits that deliver high harmonic output with minimum value of crystal current are shown above.

(point A) and the crystal. Only enough turns should be used in the feedback portion of the oscillator tank circuit to insure that the crystal will start promptly each time the oscillator is turned on. Fundamental frequency crystals operated in the overtone mode will require a relatively large amount of feedback and consequently a greater number of turns on the feedback coil than are required by specially cut overtone crystals.

When the capacitive circuit of Figure 3C is employed, feedback may be controlled by adjusting the variable capacitor between the crystal and ground. Feedback is reduced as the capacity value is increased.

A unique cathode coupled oscillator is shown in Figure 3D that is capable of output as high as the ninth crystal overtone. The crystal is coupled between the cathodes of a twin triode with the left triode section acting as a grounded grid oscillator, and the right section acting as a frequency multiplier.

VHF Tank Circuits

Certain tubes having high output capacitance are poor performers in the 144 mc and 220 mc bands, as the tank inductance tends to disappear into the internal leads of the tube. Good results may be obtained with such high capacitance tubes as the 2E26 and the 6146 in the VHF region by the use of the so-called *"series tuned" tank circuit,* shown in Figure 4A, wherein the output capacity of the tube and the tuning capacitor combine to form a split tank circuit. In actuality, the configuration is a conventional parallel tuned circuit, as shown in Figure 4B.

Linear Tank Circuits

In VHF circuits, not only must the tube losses be kept very low, but the losses in the associated circuits must also be kept as low as possible. In

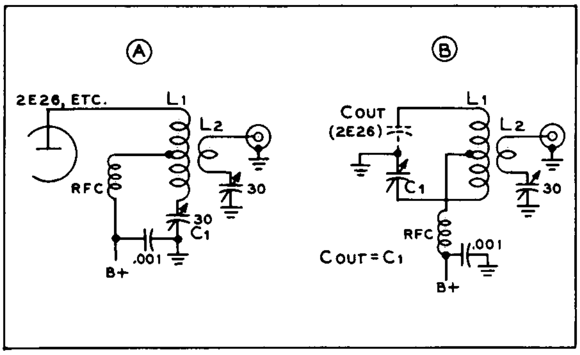


Fig. 4 "Series tuned" VHF tank circuit is shown as parallel circuit, with tube output capacity forming half of split tank circuit.

other words, the circuits when unloaded must be of the highest possible Q. For this reason, the tuned circuits associated with VHF equipment are often resonant sections of transmission line rather than coil and capacitor combinations (Figure 5). The Q of a line section can be made much higher than that of a conventional tank circuit because it is feasible to make a tuned line of conductors of larger diameter than is possible with the conventional inductor, making the skin effect less. In addition to their low losses, tuned lines are used as circuit elements because the tube leads may act as extensions of the transmission line. Thus the interelectrode capacitances and lead inductances are incorporated as part of the tuned circuit. Linear tank circuits may be either one quarter wavelength long as shown in Figure 5A, or one half wavelength long, as shown in Figure 5B. For highest efficiency, the lines should be silver plated to reduce r-f skin resistance, and should be inclosed in an r-f tight shield to reduce radiation losses.

The linear tank circuit may be combined with the usual coil-capacitor circuit to form a two band tank assembly, as shown in Figure 5C. Such a configuration is often used for operation on the 50 mc and 144 mc bands without the necessity of changing coils or adjusting sliding bars on the linear lines. The 50 mc inductance is placed at a low potential point on the 144 mc one-half wavelength linear tank circuit. Tuning capacitor C1 is placed across the free end of the linear system resonating it to 144 mc, while at the same time permitting 50 mc resonance in the circuit L1-C1.

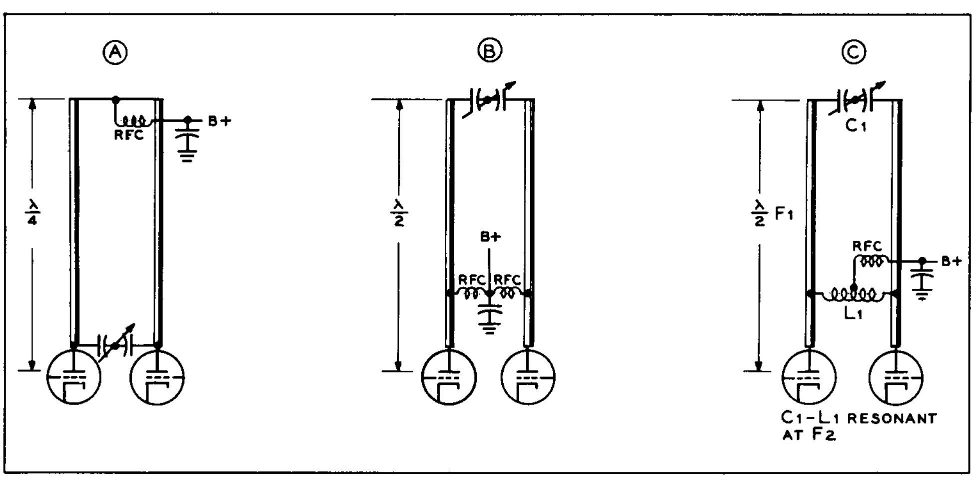


Fig. 5 Three types of VHF linear tank circuits that are discussed in the text.

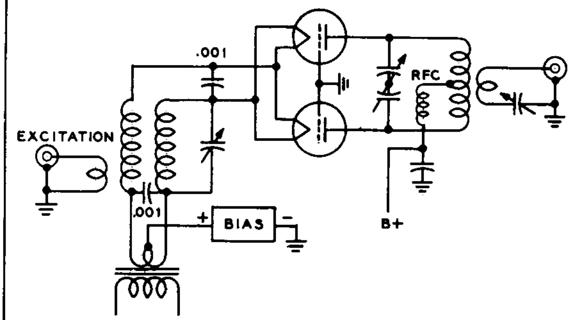


Fig. 6 Grounded grid VHF amplifier provides highly stable circuit with minimum of capacitive loading. No neutralization is required as grounded grid provides shield between input and output circuits.

The Grounded-Grid R-F Amplifier

One of the undesirable characteristics of the conventional triode r-f amplifier is that the stage must be neutralized to prevent self oscillation. As the frequency of operation of the stage is increased the stage becomes increasingly difficult to neutralize because of the inherent inductance in the grid and plate leads of the circuit, and the leads to the neutralizing capacitors. In addition, the extra circuit capacitance of the neutralizing capacitors further loads the plate and grid circuits of the amplifier, decreasing the maximum frequency of operation of the stage.

These difficulties may be overcome by the use of a grounded grid configuration, as shown in Figure 6. Excitation is fed to the cathode circuit of the amplifier and output taken from the plate circuit. The grids of the tubes are at r-f ground potential, and act as a shield between the input and output circuits. The output capacitance of such a stage is reduced to one-half the value obtained if the tubes were operated in a conventional grounded cathode circuit. In addition, the feedback path between output and input circuits is greatly reduced, since the plate to cathode capacitance of the normal VHF triode tube is very low.

In a conventional grounded grid amplifier, the input and output circuits may be considered to be in series. A large amount of excitation is required, but a proportion of this energy appears in the plate circuit of the stage as useful output. The usual power gain for a grounded grid amplifier runs between three and ten, depending upon the chosen operating parameters. In addition, the grounded grid amplifier cannot be modulated 100 per cent unless the exciting stage is also modulated. A modulation factor of about 70% is required for the exciter to permit 100 per cent modulation of the grounded grid amplifier.

It is important to note that full excitation power cannot be applied to the grounded grid stage until the plate circuit is loaded, otherwise the grids of the amplifier tubes would be damaged from excessive excitation. Excitation must be built up gradually as the grounded grid amplifier is loaded to the antenna circuit.

In spite of these minor difficulties, the grounded grid stage is becoming more popular for VHF work, especially in high power circuits, such as those required for moon reflection work.

CHAPTER XI

VHF Transmitter Construction

This chapter covers in detail the construction, alignment and testing of various types of VHF transmitters, ranging from a small portable-mobile 50 mc and 144 mc unit, to a high power amplifier suitable for moon reflection transmissions.

For convenience, a nomenclature reference for the more important components shown in the schematics is given in Figure 1. These designations are followed in all the circuit diagrams. In addition, test voltages and currents are given in the schematics for the vital measurements for proper transmitter operation.

A Low-drain Transmitter for 144 mc.

This compact, fool-proof transmitter is designed to work in conjunction with a 300 volt, 100 milliampere power supply. It may be used for mobile, or Civil Defense work, or for home station use. It employs only four tubes, and delivers a fully modulated carrier of 2 watts on 144 mc. This circuit is well suited for the Novice operator as it is inexpensive, and trouble free.

Circuit Description

The circuit of the low-drain transmitter is shown in Figure 3. An 8.1 mc crystal is used in an overtone oscillator stage, delivering 24.3 mc energy. A 12AV7 VHF double triode is employed in this circuit, the first section functioning as the oscillator, and the second section as a frequency tripler to 72.9 mc. A transmitting-type 5763 acts as a frequency doubler from 72.9 mc to 145.8 mc. The plate circuit of this stage is series tuned for maximum efficiency. The coaxial antenna circuit is link coupled to the doubler tank coil, and series tuned by capacitor C4.

FIGURE 1. SEE NOMENCLATURE CHART OF CHAPTER 8 IN REFERENCE TO SCHEMATICS.

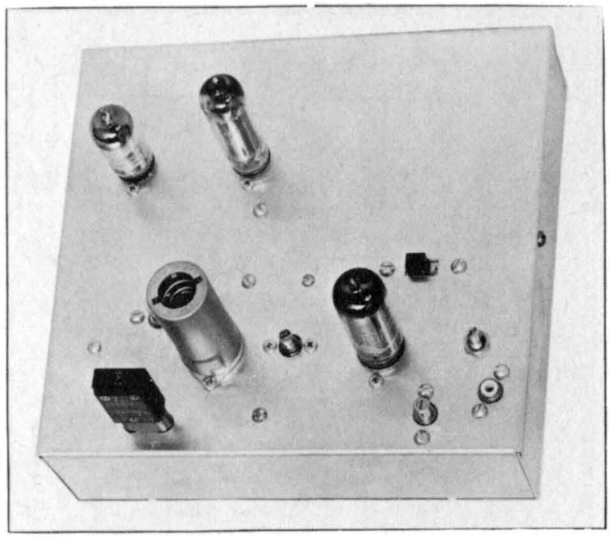


Fig. 2 Miniature 144 mc transmitter is well suited for mobile or Civil Defense work. Running 7 watts input, total plate current drain is only 100 ma at 300 volts. Entire transmitter may be run from single vibrator supply for mobile or portable operation.

A tune-up switch (S1) is incorporated in the screen circuit of the 5763, permitting the screen voltage to be removed from the tube during the adjustment process. All of the r-f circuit leads are suitably bypassed and filtered to reduce harmonics and spurious emissions to an absolute minimum.

The doubler-amplifier stage is plate and screen modulated by a 6AQ5 class-A pentode. The modulator is choke coupled to the 5763 by means of a tapped audio choke that permits the correct impedance match between the two stages. Modulation in excess of 90% is afforded by this step-down ratio. A 6AB4 high-mu triode amplifier stage permits the use of an inexpensive crystal microphone with this transmitter. Modulation level may be adjusted by varying the distance between the lips and the microphone, eliminating the need of a variable gain control.

For 12 volt mobile operation, a 6417 may be substituted for the 5763, and a 12AQ5 for the 6AQ5. The filaments of the 12AV7 may be wired in series, and a 42 ohm 2 watt resistor placed in series with the filament of the 6AB4.

Transmitter Construction

The transmitter is constructed upon an aluminum box-chassis measuring 7½″ x 7″ x 2″ *(L.M.B. #20)*. The layout of the major components may be seen in the top view photograph, Figure 2. The two r-f stages are placed across the front of the chassis, with the 12AV7 crystal oscillator stage to the left. The 5763 r-f amplifier-doubler is to the right, with the screen switch, S1, directly behind it. At the rear of the chassis are the audio stages, with the 6AQ5 to the right.

The under-chassis configuration is shown in Figure 4. Center-to-center spacing of the r-f tube sockets is 2¾″, and the sockets are placed 1¾″ from the front edge of the chassis. The audio sockets are spaced 2″ apart, center-to-center, and are located 2″ from the rear edge of the chassis. The modulation choke T1 is bolted to the rear lip of the chassis with the power plug P1 immediately to the right of it.

All filament wiring should be done first, and all required grounds made at each tube socket. Short, direct grounding leads from the socket pins to the grounding "ears" of the sockets should be used. Ground the rotors of

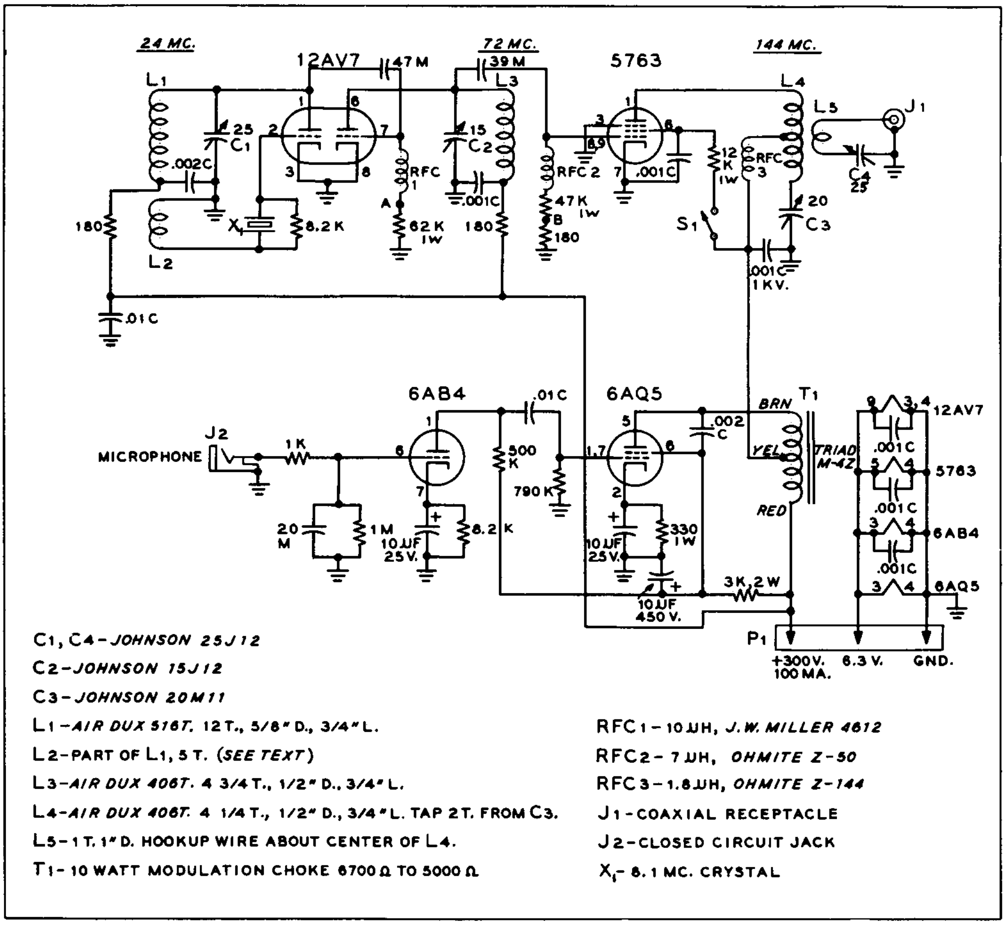


Fig. 3 Overtone oscillator using 8 mc crystals provides 72 mc output to drive miniature 5763 pentode as plate modulated doubler to 144 mc band.

all tuning capacitors. Coil L1-L2 is made from a single section of miniature air inductor. A center turn is cut to make two coils, but the polystyrene support rods should not be cut. Separation between L1 and L2 is therefore only the distance of the removed coil turn. One end of L1-L2 is supported by the stator rod of C1, and the other end of the double inductor by one pin of crystal socket X1. A .002 ufd ceramic capacitor is connected between the two inner coil leads. The inner coil lead of the grid section is grounded to an "ear" of the 12AV7 socket, and the lead from the plate coil is attached to the 180 ohm plate decoupling resistor.

RFC-1 and the 62K grid resistor of the second section of the 12AV7 are supported at *point A* by means of an insulated terminal. This junction is used to check the grid excitation of the tripler stage, as explained later. The 72 mc plate circuit of the tripler stage is located between the two r-f tube sockets. Coil L3 is supported at one end by the stator of capacitor C2. RFC-2 is attached to pin 8 of the 5763 socket, and the junction between this component and the 47K grid resistor is supported by one end of a two terminal phenolic tie-strip. The other terminal of the tie-strip forms *point B,* the grid excitation check point of the 5763 stage.

Plate coil L4 of the 5763 tube is mounted between pin 1 of the r-f amplifier-doubler socket and the stator of C3. The center-to-center spacing

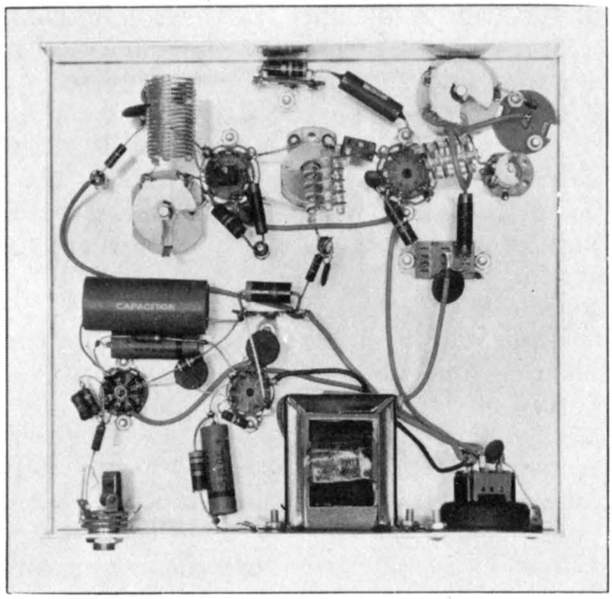


Fig. 4 Bottom view of 5763 transmitter. The 12AV7 circuit is at upper left, with 5763 circuit at right. The modulation choke is located on chassis wall between microphone jack and power plug. Audio tube sockets are at lower left of chassis.

between the socket and C3 is 1½″. RFC-3 is tapped on L4 and the opposite end of this choke is supported by a terminal of S1. The .001 ufd, 1 KV plate bypass capacitor is attached at this point.

Layout of the audio section is quite straightforward. Components are mounted between tube socket pins wherever possible. A two terminal phenolic tie-strip is mounted near the center of the chassis to support the 3K, 2W dropping resistor and the positive terminal of the 10 ufd, 450 volt filter capacitor. The negative terminal of the filter capacitor is grounded to an "ear" of the 6AB4 tube socket.

When the transmitter wiring has been completed, all circuits should be checked with an ohmmeter for shorts, transpositions, opens and accidental grounds.

Transmitter Adjustment

Upon completion and check-out, the transmitter may be connected to a 300 volt, 100 milliampere power supply. An 8 mc crystal, the 12AV7, and the 5763 tubes should be placed in their sockets. Tune-up switch S1 should be opened, removing screen voltage from the amplifier-doubler stage. If a grid-dip oscillator is at hand, the tuned circuits may be set to their approximate resonant frequencies, as indicated in Figure 3. Plate power should be applied to the transmitter through a 0-100 d.c. milliameter. The oscillator plate current should be about 30 ma, and should drop as capacitor C1 is varied, indicating crystal oscillation. The degree of oscillator feedback is controlled by the number of turns on coil L2. For the usual 8 mc crystal, five turns is correct. With an unusually active crystal, or a "24 mc" overtone crystal, the turns on L2 may have to be reduced to perhaps only one or two. A small pilot lamp attached to a one turn loop and held near L1 will indicate oscillation. If the bulb shows an indication over most of the range of C1 it is a sign that the stage is self-oscillating, and the number of turns on L2 should be reduced, as explained in chapter 10 of this Handbook.

A high resistance voltmeter attached to check point A through a 2.5 millihenry r-f choke should indicate a developed bias of -90 volts or so.

A 0-10 d.c. millammeter is next connected between point B and ground, and capacitor C2 tuned for maximum grid current of the 5763 stage. About 1 ma of grid current should be obtained.

A No. 46 (blue bead) pilot lamp should now be connected to the terminals of J1, and S1 closed. Capacitor C3 is tuned for resonance, indicated by a slight kick of the measuring meter in the B-plus lead to the transmitter. Total transmitter plate current should be approximately 45 ma. As C4 is adjusted the lamp should light brilliantly. Coupling between L4 and L5 should be adjusted for maximum brilliance, while holding the plate current to 50 ma, or less.

The audio tubes are now placed in their sockets, and the total transmitter plate current will rise to about 95 ma. A microphone is plugged into J2. Under modulation, the plate current of the transmitter will kick to about 97 ma, and the dummy antenna bulb should increase in brilliance. If the coupling between L4 and L5 is too tight, the lamp will dim under modulation. When these adjustments have been completed, the transmitter may be attached to an antenna, and the 5763 stage retuned for optimum output. Either a vibrator power supply or an a.c. operated supply may be employed to power the transmitter.

100 Watts on 50 Mc—The "Simple Sixer"

This medium power transmitter is the ultimate in simplicity for six meters. Employing only two tubes, it is capable of delivering a 60 watt, trouble-free carrier, with a minimum of spurious emissions. The transmitter is designed to be operated from a dual plate supply delivering 300 and 500 volts. It may be modulated by a pair of 6L6 tubes in class AB, or the plate and modulation power may be obtained from some of the popular 120 watt medium frequency amateur transmitters. Operated at reduced input, the transmitter is ideal for the Technician amateur interested in 50 mc operation.

Circuit Description

Only two tubes are used in the ultra-simple circuit shown in Figure 6. A 6AG7 serves as a harmonic oscillator, delivering 50 mc output from a 25 mc crystal. The use of such a high frequency crystal eliminates many of the spurious responses that a low frequency crystal may produce, with the resulting interference to television channels 2 and 3. The cathode circuit of the tri-tet oscillator is tuned to 38 mc for greatest harmonic output at 50 mc.

Inductive coupling is employed for maximum energy transfer from the oscillator stage to the 829B class-C push pull amplifier. In addition, this double tuned circuit helps to reject high order oscillator harmonics that might otherwise cause television interference. The 829B is cross-neutralized for maximum stability. Screen voltage to this stage may be removed for tune-up purposes by means of switch S2.

A small butterfly tuning capacitor (C4) is employed in the plate circuit of the 829B. The rotor of this capacitor is not grounded, and the capacitor is insulated from the chassis to prevent voltage breakdown under peaks of modulation. The antenna circuit is inductively coupled to the 829B plate tank circuit, and is series tuned by capacitor C5.

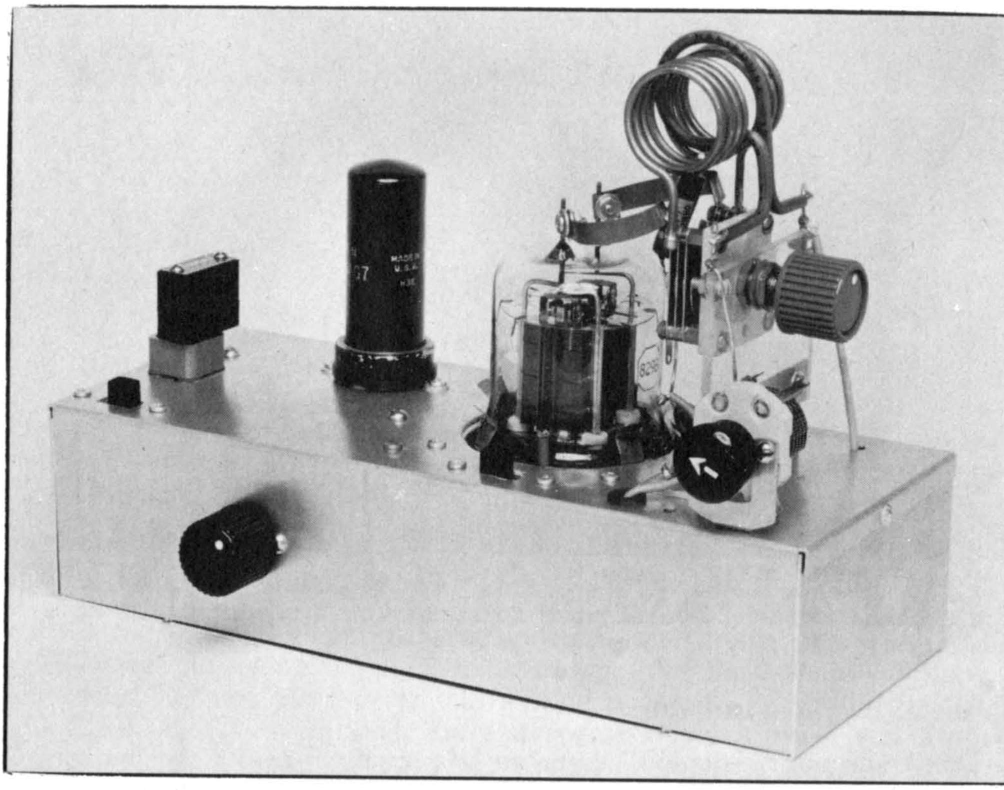


Fig. 5 The "Simple Sixer" 50 mc transmitter employs a 25 mc crystal and two tubes. Up to 100 watts input may be run by this little powerhouse. At the left are the crystal and the 6AG7 oscillator tube. The 829B amplifier and output circuit are at the right. Amplifier screen switch (S2) is in front of the amplifier tube. Oscillator tuning capacitor (C2) is located on the front of the chassis, and antenna tuning control (C5) is to right of the 829B.

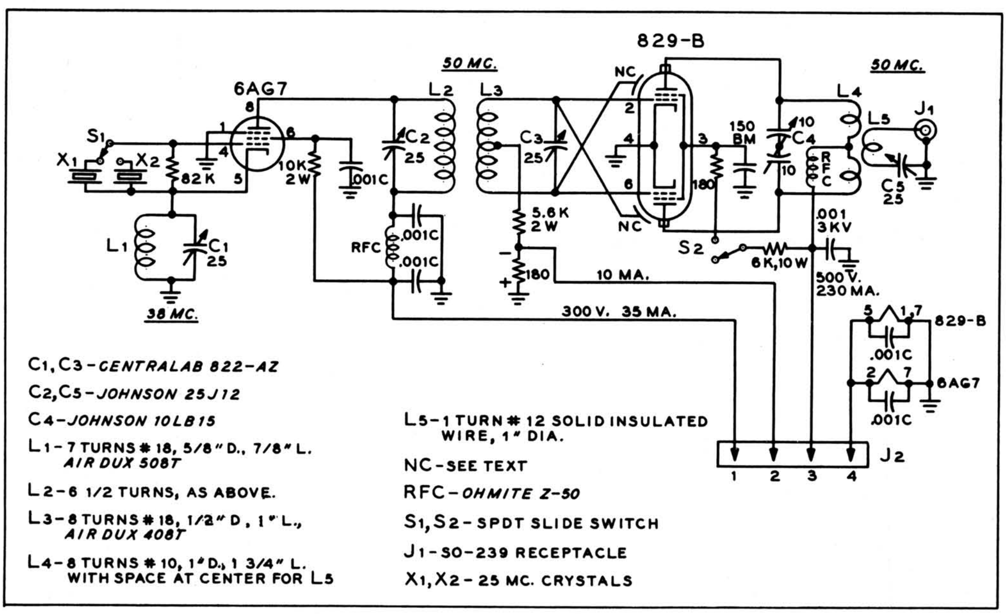


Fig. 6 Inductive coupling is used between the oscillator and amplifier stages to reduce radiation of spurious frequencies. The 829B amplifier is neutralized for maximum stability. Transmitter output is over 60 watts.

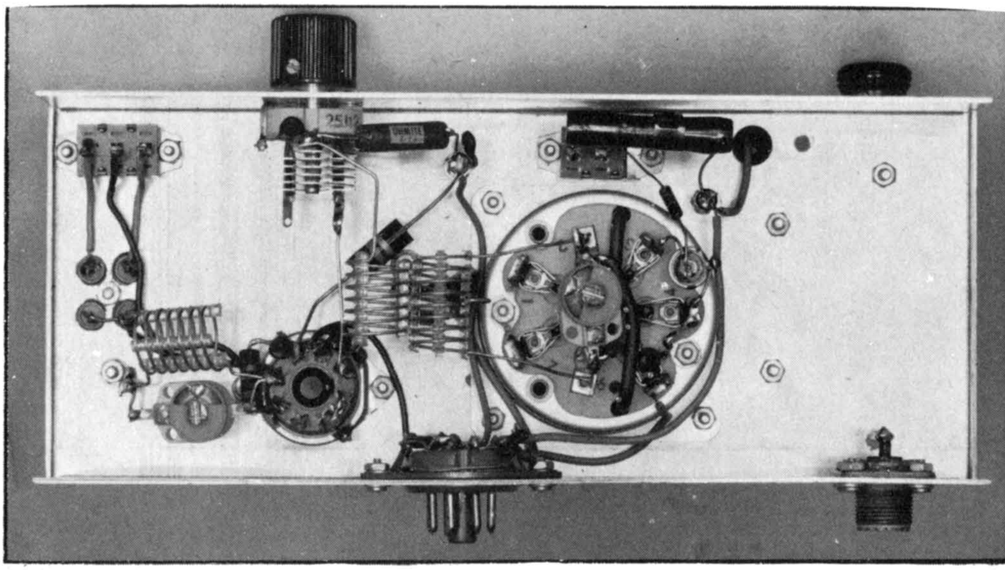


Fig. 7 Under-chassis view of "Simple Sixer." 6AG7 socket is at left, with cathode coil attached to pin 5. Double tuned interstage circuit is at center. Crossover neutralizing leads may be seen atop the 829B amplifier socket.

Transmitter Construction

The transmitter is built upon an aluminum box-chassis measuring 10" x 4" x 2½" *(L.M.B. #144)*. Placement of the major components may be seen in the top view, Figure 5. The plate circuit components of the 829B are mounted on a 2" x 2½" piece of ⅛" lucite attached to the chassis by means of a 2½" piece of aluminum angle stock.

Placement of the parts under the chassis are shown in Figure 7. The tuned cathode circuit (L1-C1) is at the left, and the coupling inductances (L2-L3) are at the center. The oscillator plate tuning capacitor C2 is mounted on the front lip of the chassis. The rotor of this capacitor is at plate potential, so the capacitor is mounted in a ½" hole to allow sufficient clearance to prevent arc-over to ground.

The 6AG7 socket is 3" from the end of the chassis and 1" from the rear edge. The 829B socket is centered along the chassis axis and 4" from the right end, as viewed from the top. Tune-up switch S1 is placed directly in front of the 829B socket. Oscillator plate coil L2 is mounted between pin 8 of the 6AG7 socket and the rotor of capacitor C2. RFC attaches to the rotor terminal of C2, as does the .001 ufd ceramic bypass capacitor. The opposite end of RFC is supported by a phenolic tie-point.

Grid coil L3 of the 829B is supported directly from pins 2 and 6 of the amplifier tube socket and lies in the same plane as L2 and about ⅛" away from it. Grid tuning capacitor C3 is also attached to the grid pins of the 829B socket. The two neutralizing capacitors (NC) for the 829B are made of short lengths of plastic covered #12 wire. These leads are criss-crossed beneath the chassis and pass through the two socket holes located adjacent to the grid pins. The leads project above top level of the chassis about ⅜"

To insure maximum stability, the screen of the 829B is bypassed to ground with a low inductance button mica capacitor mounted in the socket rivet hole adjacent to pin 4. A 180 ohm, ½-watt composition resistor is placed in the screen lead between the button mica capacitor and S2 to

isolate the screen circuit from stray r-f currents flowing in the power leads.

Filament pins 1 and 7 of the 829B socket are grounded by means of soldering lugs placed beneath the head of a bolt passed through the hollow socket rivet between these two pins. Filament pin 5 is bypassed to ground by a ceramic capacitor grounded to a bolt passed through a socket rivet.

The amplifier plate coil L4 is soldered directly to the stator arms of the butterfly capacitor C4. Spacing between the turns of the coil is varied until resonance is obtained with C4 about ⅔ meshed. The antenna pickup coil L5 is made of a length of plastic insulated #12 wire attached at each end to bolts passed through the upper corners of the lucite plate. One end of L5 is then attached to C5 whose rotor is grounded to the chassis, and a lead from the other end passes through a hole in the chassis to J1 mounted on the rear wall of the chassis.

Transmitter Adjustment

After the wiring has been checked, a 25 mc crystal should be plugged in one crystal socket, and S1 set to place that crystal in the circuit. The 6AG7 and 829B tubes should be placed in their respective sockets, and tune-up switch S2 opened. A 300 volt power supply should be connected to the 6AG7 stage through a 0-50 d.c. milliammeter. A 0-25 d.c. milliammeter is connected between pin 2 of J2 and ground to measure the grid current of the 829B. As C2 is adjusted, the plate current to the 6AG7 will drop to approximately 30 ma, and a reading should be obtained on the grid meter of the 829B stage. C2 and C3 are adjusted for maximum grid current, then C1 is peaked to maximize this reading. The spacing between L2 and L3 is not critical, but it might be varied slightly to find the optimum position. A reading of 10 to 12 milliamperes of grid current should be obtained.

Power should now be turned off, and capacitor C4 set to tune the plate circuit of the 829B to 50 mc. The turns of L4 may be spread or squeezed slightly, if necessary. A grid-dip oscillator should be used for this adjustment. Plate power should be applied to the 6AG7 and amplifier plate tuning capacitor C4 tuned through resonance. The grid meter should be observed closely during this test. If the stage is not completely neutralized, the grid current will kick as C4 is tuned. The neutralizing leads (NC) must then be varied in length until this reaction is no longer apparent. The leads should not be changed over ⅛″ of an inch at a time, and the change must be done equally to each lead.

Once the amplifier is neutralized, a 75 watt lamp bulb is attached to coaxial antenna receptacle J1, and a 500 volt power supply attached to the amplifier plate terminal, J2. The B-plus lead should pass through a 0-300 d.c. milliammeter. Plate voltage is applied, and tune-up switch S2 closed. Amplifier tuning capacitor C4 should be resonated for minimum plate current, and antenna capacitor C5 adjusted for a plate and screen current reading of 230 ma. Grid current to the 829B should be 10 ma.

If desired, a 5894 may be substituted for the 829B. The 5894 is self-neutralized in the 50 mc region, and the two neutralizing capacitors (NC) should be omitted. The screen dropping resistor should be changed to 15K, 10 watts, and the bias resistor should be changed from 5.6K to 20K, 5 watts. Under these conditions, grid current of the 5894 is 5 milliamperes, and screen current is 16 milliamperes, at a plate voltage of 500. Combined plate and screen current is 216 milliamperes. Power output is in excess of 60 watts.

A 50 Mc R-F Unit for the "Ranger" Transmitter

One of the most popular transmitters on the market today is the 75 watt Viking "Ranger." The circuitry of the Ranger permits the plate power and audio output to be applied to an auxiliary piece of equipment through a 9 pin receptacle mounted on the back of the Ranger.

Described herewith is a 60 watt r-f amplifier package that attaches to the Ranger and permits 50 mc operation utilizing the power, modulator, and control circuitry of the Ranger. No wiring changes or other alterations are necessary in the Ranger when this r-f unit is used.

Circuit Description

The Six meter r-f unit is shown in Figure 8 working in conjunction with the Ranger transmitter. If desired, it may be run with its own modulator and power supply, as described later. The circuit employs three tubes. A 6AG7 is used as a tri-tet harmonic oscillator, doubling to 50 mc from a 25 mc crystal, as shown in Figure 9. The oscillator is inductively coupled to the grid circuit of a neutralized 6146 class-C r-f amplifier. A 6AQ5 is used as a protective clamp tube in the 6146 screen circuit.

Connection between the units is made via a multi-wire cable that terminates in a 9 pin plug that matches the receptacle on the rear of the Ranger transmitter. The B-plus voltages to the 50 mc r-f unit are controlled by relay RY-1 for standby purposes. The relay coil is actuated by the antenna relay circuit of the Ranger transmitter.

R-f filter circuits are placed in all power leads to the 50 mc r-f unit to reduce the radiation of spurious harmonics. The complete transmitter is housed in one of the new TVI- suppressed aluminum cabinets to aid in the suppression of unwanted harmonics. As a final harmonic suppression measure the panel meter is shielded and the leads to it are filtered.

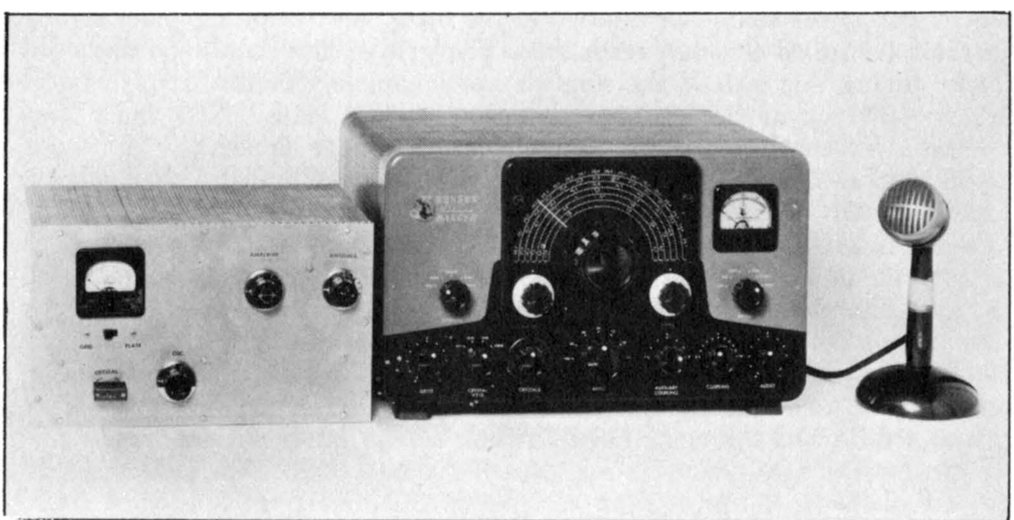
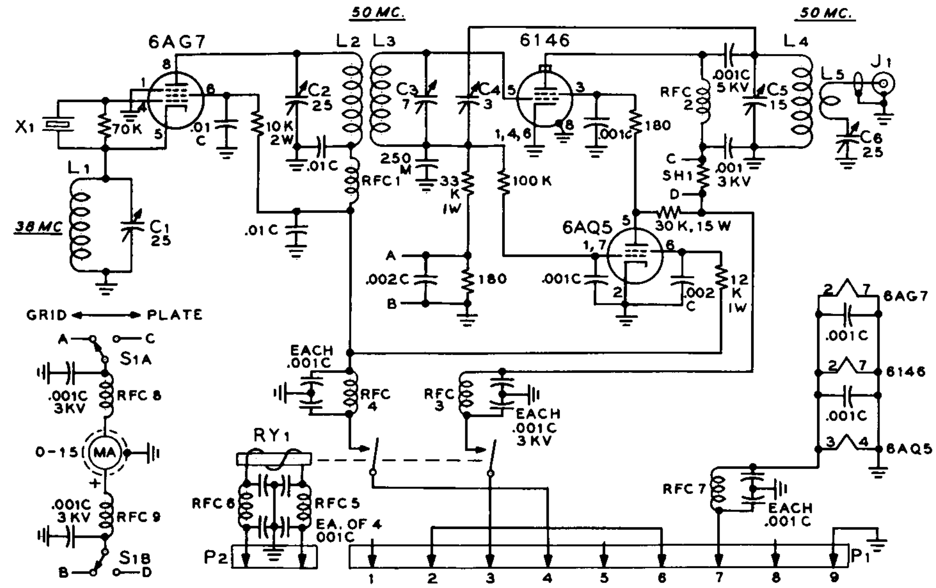


Fig. 8 50 mc operation of the Ranger transmitter is possible with the use of this simple r-f unit. Making use of the power supply and modulator of the Ranger, this 50 mc adapter runs the full 60 watt power level. Power connections between the two units are made via the nine pin receptacle on the back apron of the Ranger. No changes need be made to the Ranger transmitter.



C1,C3 - *CENTRALAB 822*
C2,C6 - *JOHNSON 25J12*
C4 - *JOHNSON 30M8* (SEE TEXT)
C5 - *BUD LC-1654*
L1 - 7 T. #18, 5/8" D., 7/8" L. *AIR-DUX 508T*
L2 - 6 1/2 TURNS, AS ABOVE
L3 - 9 TURNS, AS ABOVE
L4 - 4 TURNS #10, 1" D., 1" L.
L5 - 3 TURNS #14, 1" D., 1/4" L.

RFC1, RFC2 - *OHMITE Z-50*
RFC3-6, 8-9: *OHMITE Z-144*
RFC7 - *NATIONAL R-60*
RY1 - *ADVANCE AM2C*, 115 V. COIL
SH1 - SEE TEXT
X1 - 25 MC. CRYSTAL, *BLILEY AX-3*
P1 - 9-PRONG PLUG *86-PM9-11*
CABINET - *LMB # 1287*

Fig. 9 Schematic of Ranger r-f unit for 50 mc. Complete circuit shielding and filtering reduce TVI problems to a minimum. 6146 amplifier is bridge neutralized and employs a 6AQ5 screen clamp tube for overload protection during tune-up.

Transmitter Construction

The transmitter is housed in the new *L.M.B. #1287* TVI-suppressed aluminum cabinet measuring 12" x 8" x 7". Before the cabinet is assembled, the chassis and front panel are drilled, and the major components are mounted on the chassis. The top and under-chassis views of Figures 10 and 11 show the placement of the major components.

All socket ground connections and filament leads are wired first, and the rotors of C2 and C6 are grounded. Plate coil L2 of the oscillator stage is supported at one end by the stator of C2, and at the other end by a phenolic tie-point which also supports one end of RFC-1 and the .01 ufd ceramic plate bypass capacitor. Grid coil L3 of the 6146 stage mounts along the same axis as that of L2, and about 3/8" away. Coil L3 is supported at one end by pin 5 of the 6146 socket, and by a phenolic tie-point at the other which also supports one terminal of the 250 uufd mica bypass capacitor. The ceramic grid tuning capacitor C3 is attached across the leads of coil L3 and is supported by it.

The only critical leads in the amplifier configuration are those in the neutralizing circuit. These should be as short and as heavy as possible to have a minimum amount of inductance. If the lead inductance is excessive,

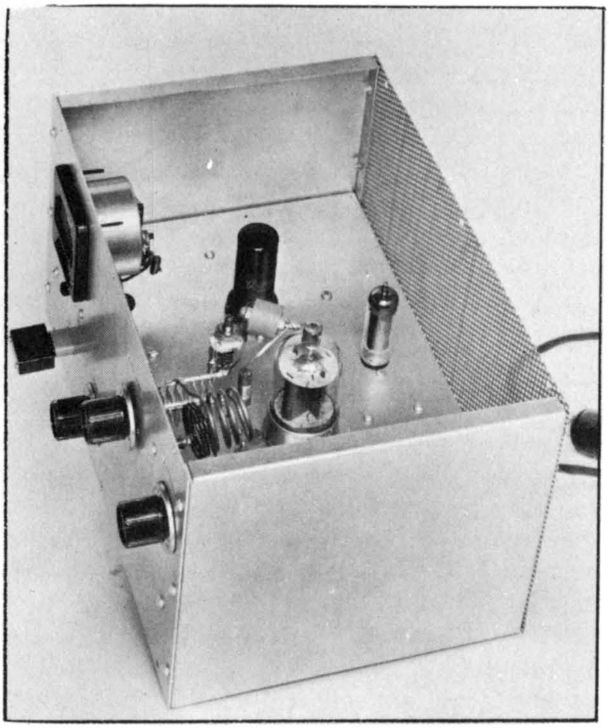


Fig. 10 The 50 mc r-f unit with top cover removed. Neutralizing capacitor C4 is attached to the 5KV ceramic plate blocking capacitor by a short length of brass strap. Plate circuit tuning capacitor C5 and the antenna resonating capacitor C6 are panel mounted in front of the 6146 amplifier tube. At rear of chassis is the 6AQ5 clamp tube.

the stage will refuse to neutralize. The 250 uufd mica capacitor is mounted vertically directly between the end of the grid coil and a soldering lug placed under the phenolic tie-point. The leads of this capacitor are less than ¼″ long. The lead from the grid coil to neutralizing capacitor C4 passes through a rubber grommet placed in a ¼″ hole in the chassis. This lead is made from a short section of #10 copper wire. Directly above the ¼″ hole the lead is soldered to the stator rod of the neutralizing capacitor, as shown in Figure 10.

To insure that the residual plate lead inductance is at a minimum, a *Centralab type 358S* capacitor is used for the plate blocking capacitor. This is a .001 ufd, 5KV unit with threaded studs. One stud is attached directly to the plate cap of the 6146, and the other stud supports the neutralizing capacitor, C4, by means of a short length of brass sheet cut from thin shim stock. A second short length of brass completes the connection between the blocking capacitor and the plate tuning capacitor, C5, which is attached to the front panel.

Neutralizing capacitor C4 is quite small in capacitance, but requires a greater spacing between the plates than is commonly available in such a small unit. Accordingly, a *Johnson30M8* capacitor is modified by removing ⅔ of the plates, reducing the capacity and increasing the voltage breakdown. Every fourth plate is left in the capacitor, the intervening two plates being removed by twisting them gently with needle-point pliers. When completed, 5 plates will remain in both the rotor and stator, and the spacing will be tripled. If an error is made in removing plates, they may be soldered back in place with the aid of a very small iron. Care must be taken to see that the spacing is increased in a symmetrical manner, and that the rotor and the stator configuration are identical.

The plate choke (RFC-2) of the 6146 is mounted vertically between the

plate cap of the tube and a ceramic feed-thru insulator placed on the chassis. The bottom end of this choke is bypassed above the chassis by a .001 ufd, 3KV ceramic capacitor. Plate coil L4 is attached directly to the stator rod of C5, and to the center grounding lug of the variable capacitor.

The antenna link, L5, is attached at one end to the stator of C6, and at the other to a phenolic tie-point mounted to the front panel. A short length of small coaxial line (RG-59/U) is attached to the coaxial antenna receptacle and terminates at this tie-point.

The meter leads pass through a grommet mounted in a ½″ hole drilled in the chassis. Meter switch S1 is mounted directly above this hole, and the 0-15 d.c. milliammeter is placed directly above the switch. When S1 is thrown to the grid position, the meter is connected across a 180 ohm resistor in the grid circuit of the 6146. The internal resistance of the meter is 6.6 ohms, so the grid resistor has no appreciable effect upon the meter reading. When the switch S1 is thrown to the plate circuit position, a "times-ten" multiplier is placed across the meter, increasing the scale reading to 150 milliamperes. The meter shunt (SH1) in the plate circuit of the 6146 stage has a resistance of about .66 ohms. An *IRC type BW* 0.62 ohm resistor may be used for SH1, or the shunt may be wound out of a suitable length of resistance wire, if desired. When completed, the reading of the

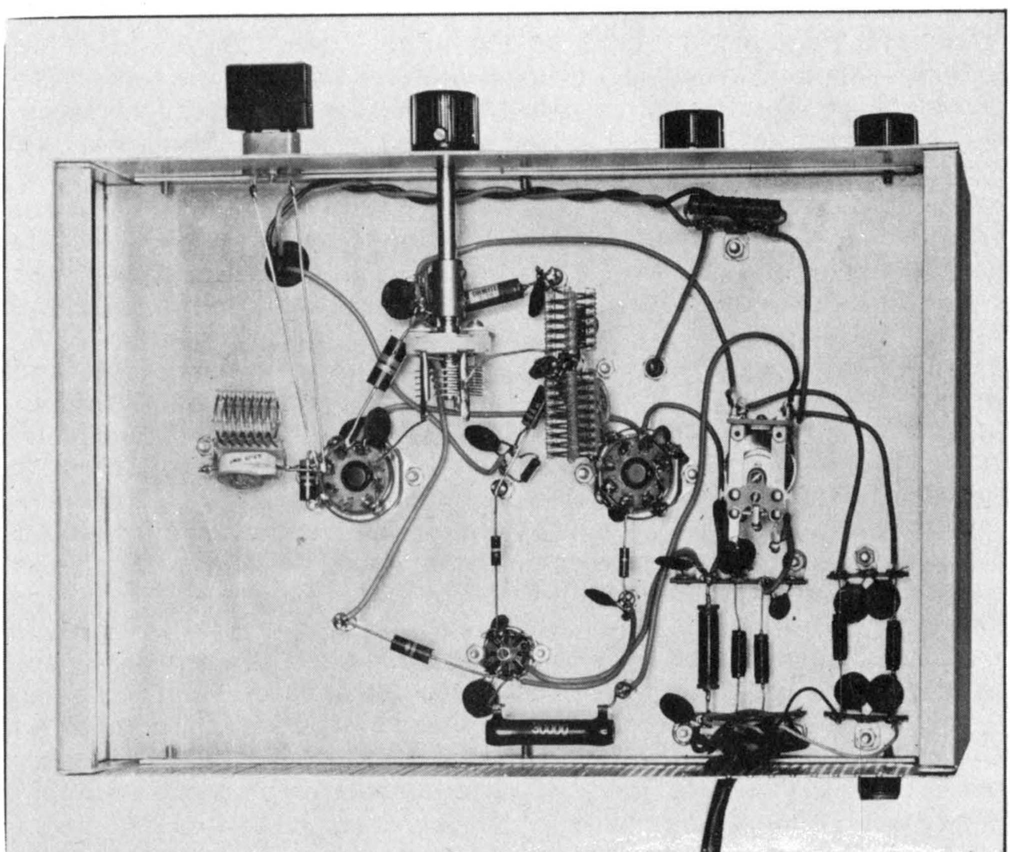


Fig. 11 Under-chassis view of 50 mc r-f unit. Double tuned circuits are at center of chassis. C2 is located between r-f tube sockets, and C3 is placed at side of coil L3. Plate circuit relay RY-1 is to right of the 6146 socket. Cathode coil of oscillator is at extreme left, and switch SH-1 at top right.

meter may be compared with the reading of the meter in the Ranger transmitter, as they should both give the same indication in the "plate" position.

A meter shield is made from the end of a tin can of the correct diameter, slid over the back case of the meter. The edges of the can are fluted to make a good ground to the back of the panel. The meter studs pass through rubber grommets placed in holes drilled in the end of the tin can, as described earlier in this Handbook. RFC-8 and RFC-9 and their associated bypass capacitors are mounted on a two terminal phenolic tie-strip attached to the back of the meter shield.

When the transmitter wiring is completed, all connections should be checked for shorts, opens, transpositions and accidental grounds.

Transmitter Adjustment

All tubes should be inserted in their sockets, and a 25 mc crystal plugged in the crystal receptacle. Capacitor C4 should be set at about mid-scale, and the 50 mc plate circuit of the 6AG7 tube resonated to frequency with the aid of a grid-dip oscillator. The plate circuit of the 6146 amplifier should also be checked with the grid-dip oscillator. The turns of L4 may be expanded or compressed to make resonance occur with C4 at half capacitance, and C5 at about ⅔ capacitance.

The lead from pin 3 of P1 to the arm of relay RY-1 should be temporarily opened, removing plate and screen voltage from the 6146 amplifier stage. The transmitter cable is now plugged in the 9-pin receptacle on the rear apron of the Ranger transmitter, and the leads from P2 are plugged in the antenna relay power receptacle of the Ranger. The Ranger control switch should be turned to the "tune" position. The filament voltage at the 6146 socket of the 50 mc transmitter is now checked. If it is below 6.2 volts, heavier leads are required for the filament circuit in the power cable. The control switch on the Ranger is next advanced to the "c.w." position. Relay RY-1 should close, applying plate voltage to the 6AG7 oscillator stage. Switch S1 should be set to the grid position, and C2 tuned for maximum grid current. C1 should then be tuned to maximize this reading. A grid current reading of about 4 milliamperes should be obtained when C2 and C3 are tuned to resonance. The spacing between L2 and L3 may be adjusted for maximum grid current. After these adjustments are made, C2 should be retuned to drop the grid current to about 2.5 milliamperes, which is the correct operating value. Do not let the grid current of the 6146 tube exceed this value for prolonged periods, or damage to the tube may result.

The plate circuit of the 6146 should now be tuned through resonance, and the grid current reading observed for a fluctuation. Any slight kick of grid current indicates incomplete neutralization. Capacitor C4 is tuned so as to eliminate this variation. If neutralization tends to occur at the minimum capacity setting of C4, the 250 uufd mica capacitor in the grid circuit of the 6146 should be increased to 300 uufd. If best neutralization occurs with C4 at full setting, the mica capacitor should be decreased in value to 200 uufd.

When C4 is correctly adjusted, no change in grid current will be noticed when C5 is tuned through resonance. It should be noted that any change in C4 will cause a slight change in the resonance setting of C5, since

the plate-ground capacity of the circuit changes with the variation of C4.

The connection between pin 3 of P1 and the arm of RY-1 may now be made, and the Ranger control switch turned to "c.w." Plate tuning capacitor C5 should be tuned for resonance as indicated on the panel meter, and antenna tuning capacitor C6 and the antenna coupling adjusted for an indicated plate current of 120 milliamperes to the 6146. Grid current should be readjusted to 2.5 milliamperes. The Ranger transmitter may now be switched to the "phone" position, modulating the 50 mc transmitter unit. The r-f circuits of the Ranger are inoperative when the 50 mc unit is in use.

A 6146 Transmitter for 2 Meters

One of the drawbacks to the proper operation of a single ended amplifier in the VHF region is the difficulty of controlling the various r-f ground return paths in the stage. The technique of proper bypassing and filtering of power leads becomes of the utmost importance, and the length of ground return leads is a significant factor in the excellence of performance of the equipment. The natural physical balance to ground of a push-pull amplifier alleviates the ground return problem to some extent, and for the beginner, the push-pull amplifier configuration is usually the easiest stage to place in proper operation.

Under proper circumstances, the 6146 beam pentode tube will perform

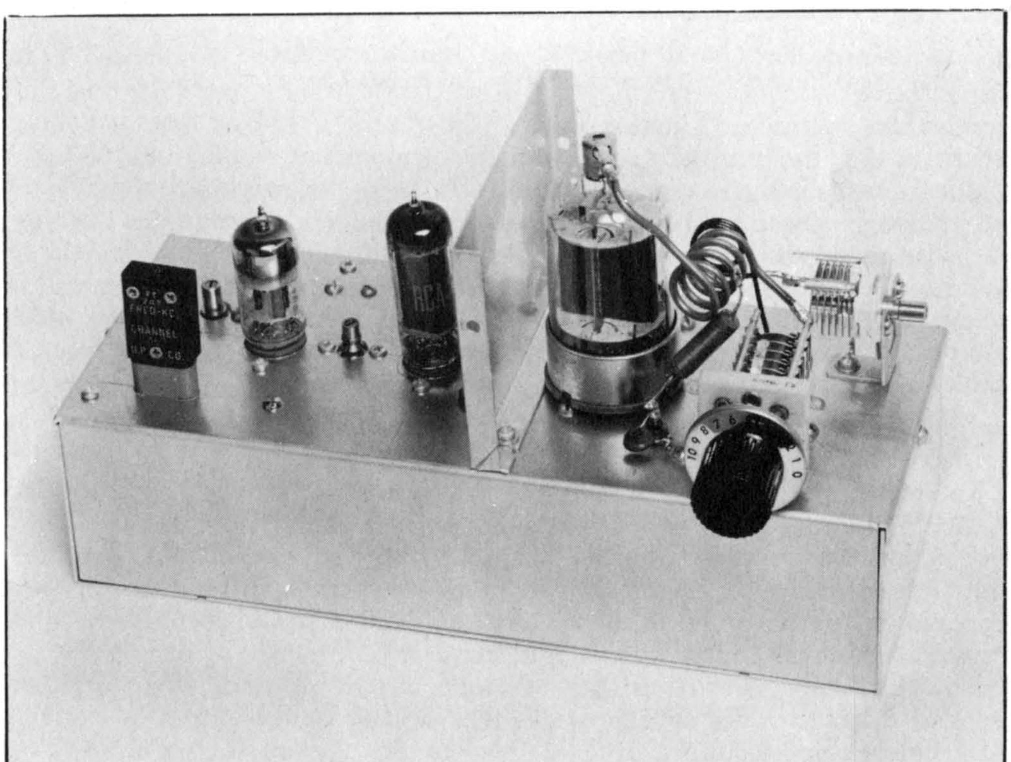


Fig. 12 Three tube 144 mc transmitter runs 40 watts input to neutralized 6146 amplifier stage. 12AV7 overtone oscillator and 8 mc crystal are at the left, with C1 and C2 projecting through the chassis. At center is the 5763 doubler stage and chassis shield. To the right is 6146 amplifier and the "series tuned" plate tank circuit. Antenna loading capacitor (C5) is at right.

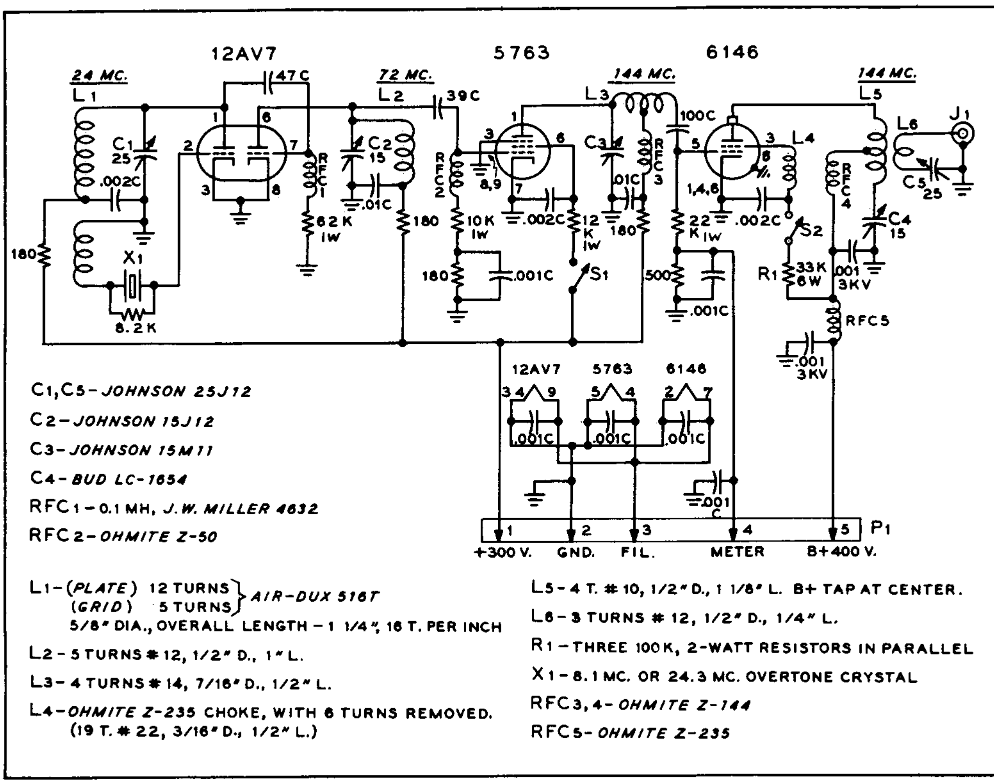


Fig. 13 Schematic of three tube 144 mc transmitter. Screen neutralized 6146 amplifier runs 40 watts input, and may be modulated by 6L6's class AB.

as a stable, neutralized single-ended amplifier on the 144 mc band, making attractive the possibility of a small, compact transmitter employing this tube. The self-neutralizing frequency of the 6146 is in the neighborhood of 100 mc, making some form of screen neutralization mandatory in the vicinity of the 2 meter band. When correctly neutralized, the 6146 will provide stable operation and freedom from parasitic oscillations that so often plague this tube when used in the VHF region. Shown in Figure 12 is a three tube 144 mc transmitter employing a 6146 amplifier running 40 watts input in a stable and fool-proof circuit.

Circuit Description

The schematic of the 6146 2 meter transmitter is shown in Figure 13. A 12AV7 high frequency double triode is used as an overtone oscillator, providing 24 mc output from an 8 mc crystal. The second section of the 12AV7 is a tripler to 72 mc. A 5763 miniature transmitting-type pentode serves as a doubler from 72 mc to 144 mc, driving the 6146 beam tube as a neutralized class C amplifier at this frequency.

A 12AV7 tube is used in the overtone circuit in preference to other types of double triodes as it provides the greatest power output at a given plate potential of all tubes tried in this circuit. The circuit parameters of this stage, and of the 5763 doubler were chosen to provide the greatest amount of grid excitation to the 6146 stage. With no plate voltage on the 6146, grid current of the amplifier is over 3.5 milliamperes.

The problem of providing a good impedance match from the plate of the 5763 doubler to the grid of the 6146 is a difficult one to solve. When

the 6146 is in an unstable condition it is highly regenerative and relatively easy to drive. When correctly neutralized, the 6146 places a low impedance load on the exciting stage, reducing the efficiency of the driver by a considerable amount. This problem was solved by the use of a split tank circuit having each tube placed at the opposite end of the coil. If the mutual coupling between the ends of the coil is great enough, sufficient excitation will be delivered to the 6146 without excessive loading of the 5763 driver stage. The plate coil (L3) of the driver stage is wound from #12 wire, with the turns very closely spaced. This configuration permits good efficiency and optimum interstage coupling.

A switch (S1) is placed in the screen circuit of the 5763 stage for tuneup purposes, permitting adjustments to be made to the overtone oscillator without damage to the doubler tube. A similar switch (S2) serves to protect the 6146 tube during driver adjustments.

The 6146 stage is neutralized by placing a small inductance (L4) in series with the screen lead of the tube directly at the socket. The turns of this coil are adjusted to provide complete neutralization of the stage, as explained later in the text.

A split tank configuration is used in the plate circuit of the 6146 amplifier to obtain highest efficiency. With a plate input of 40 watts, a measured output of over 20 watts is obtained from this transmitter.

Transmitter Construction

The 6146 transmitter is constructed upon an aluminum chassis box measuring 10″ x 4″ x 2½″ *(L.M.B. #144)*. Placement of the major components may be seen in the top and under-chassis photographs. Oscillator plate coil L1 is made from a single piece of coil stock, with one turn broken to divide the coil into two separate sections. The .002 ufd ceramic plate bypass capacitor is soldered between the coil leads directly at the coil, which is mounted between one pin of the crystal socket and a stator arm of C1. The plate coil of the tripler stage (L2) is mounted between the stator arm of C2 and a single terminal phenolic tie point which supports the .01 ufd plate bypass capacitor and the 180 ohm, ½-watt decoupling resistor.

The tuning capacitor (C3) of the 5763 stage is placed immediately adjacent to pin 1 of the tube socket. Plate coil L3 mounts between the stator of this capacitor and a 100 uufd ceramic capacitor soldered to grid pin 5 of the 6146 tube socket. RFC-3 is soldered to a turn of L3, and is supported at the opposite end by a phenolic tie point which also holds one terminal of the .01 ufd ceramic decoupling capacitor and the 180 ohm decoupling resistor.

Neutralizing coil L4 is made from an *Ohmite Z-235* r-f choke. It is soldered to pin 3 (screen) of the 6146 socket, and the opposite end is attached to a .002 ufd ceramic capacitor which is grounded to pin 8 of the 6146 socket. Pins 1, 2, 4, 6, and 8 of this socket are grounded to the retaining ring by bending them down and soldering the end of the pin to the ring. The .001 ufd ceramic filament bypass capacitor for this socket is mounted between pins 7 and 2 by the shortest possible leads.

Plate tuning capacitor C4 of the 6146 is mounted above chassis by means of the tapped mounting foot. The amplifier plate coil (L5) is supported at one end by the plate cap of the 6146, and at the other end by the stator of

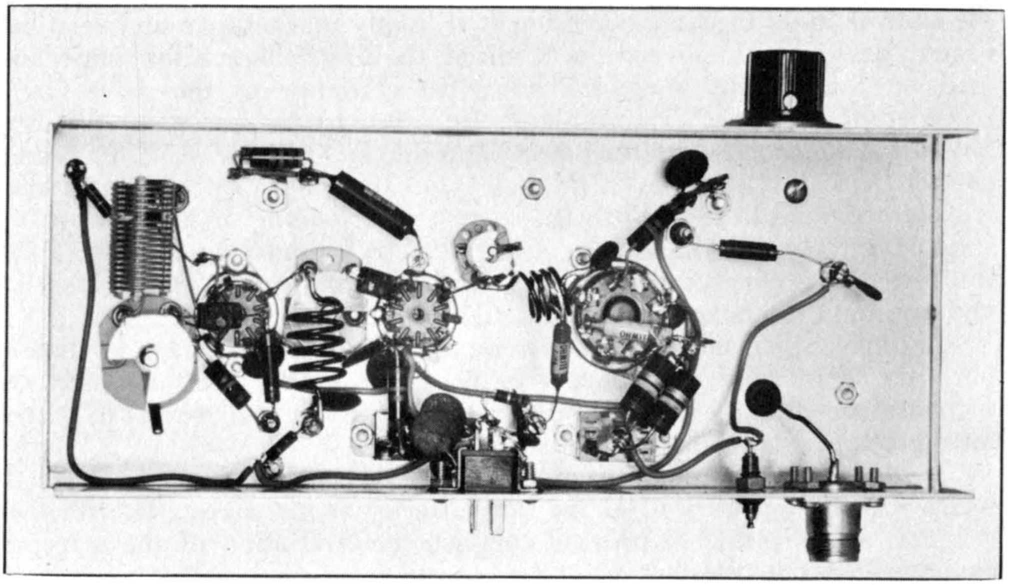


Fig. 14 Under-chassis view of 144 mc transmitter. Oscillator stage is at the left, and 5763 multiplier socket at center of chassis. The coupling coil L3 may be seen to left of 6146 socket. Coil L4 is mounted across 6146 tube socket.

C4. RFC-4 is attached to the center of L5. High voltage passes through a miniature feedthru insulator which is bypassed above the chassis by a .001 ufd, 1.5 KV ceramic capacitor.

Transmitter Adjustment

When construction is completed, all wiring should be checked for shorts, opens, transpositions, and circuit grounds. Make sure that the rotors of C1, C2, C3, C4 and C5 are grounded. Switches S1 and S2 are opened and 300 volts is applied to the exciter stages through a 0-100 d.c. milliammeter. C1 should be tuned for oscillation, indicated by a pronounced dip in oscillator plate current. Capacitor C2 should be tuned to the approximate harmonic frequency with the aid of a grid dip oscillator, and a 0-10 d.c. milliammeter placed in the grid return circuit of the 6146 stage. The 6146 should be placed in its socket, and plate voltage applied to the exciter. Switch S1 is closed and capacitors C1 and C2 are tuned for maximum grid current to the 6146. It may be necessary to spread or compress the turns of L3 to obtain maximum grid current. At least 3 milliamperes of grid current should be observed. Exciter plate current will be approximately 95 milliamperes for the two tubes.

Coil L4 should be adjusted next, as the r-f choke used has too many turns for most stable operation of the 6146 stage. A dummy load should be attached to the coaxial fitting, J1. A 50 watt lamp bulb will be satisfactory. Plate voltage is applied to the 6146, and S2 is closed. Tuning capacitor C4 should be tuned for resonance, and C5 adjusted to provide a resonance plate current of 100 milliamperes. The plate voltage to the transmitter should now be turned off, S1 opened, and the 12AV7 oscillator tube removed. Plate voltage is turned on, and the tuning capacitor of the 6146 (C4) quickly rotated through its range. The plate current of the 6146 should remain rock steady, in the vicinity of 200 milliamperes. Do not allow

the plate voltage to be on for more than two or three seconds during this test. During this test, grid current to the 6146 should be zero. If there is a tendency for the plate current to drop when C4 is rotated, or if any indication of grid current is noticed, a turn should be removed from L4 and the process repeated. With this particular circuit configuration, six turns were removed from L4 to obtain perfect stability during this test procedure.

After the 6146 is neutralized, the exciter should be placed in operation, and the antenna load adjusted for a resonance plate current of 110 milliamperes at a plate potential of 400 volts. Grid current should be 2 milliamperes, and total exciter plate current should be 95 milliamperes. It should be noted that doubler plate capacitor C3 must be tuned to a lower capacity setting when plate voltage is applied to the 6146 and the amplifier plate circuit is resonant. C3 should be readjusted for maximum grid current after the amplifier stage is loaded and tuned. This detuning effect causes no instability in the operation of the unit, and is thought to be caused by the degenerative effect and transit time loading of the particular amplifier circuit used.

The transmitter may be plate and screen modulated by 6L6 tubes operating class AB, running at a plate potential of 400 volts.

An Exciter-Transmitter for 220 Mc

Shown in Figures 15 to 18 is a two unit transmitter for the 220 mc amateur band. The exciter may be employed as a 15 watt transmitter by itself, or it may be used in conjunction with a 5894 push-pull tetrode amplifier to form a complete 80 watt unit. The exciter section makes an excellent transmitter for the Technician class amateur, and the amplifier may be added at a later date when an increase in power is desired.

The 220 mc band is in the borderline region wherein the simple coil-capacitor tank circuits used at lower frequencies begin to be replaced by linear tank circuits to obtain highest operating efficiency. At the same time, the shielding of the transmitter becomes an important part of the circuit as it tends to curb power lost by radiation from the circuit elements. At 220 mc, great attention must be paid to the power wiring of the transmitter to make sure that unwanted resonances in filament or plate leads do not rob the various stages of excitation, or create instability in the amplifier section. These design problems have been solved in this simple yet highly effective transmitter.

Circuit Description

The schematic of the 220 mc exciter is illustrated in Figure 16. Four tubes are used to reach 220 mc from an 8.14 mc crystal. A 6CL6 pentode is used in a regenerative oscillator circuit with the plate tuned to 24.44 mc (C2-L1). A second 6CL6 acts as a tripler to 73.33 mc (C3-L2). A 6360 dual tetrode is employed as a push-pull frequency tripler to 220 mc (C5-L3). The output of this stage is inductively coupled to a second 6360 tube operating as a push-pull class C amplifier at 220 mc. This amplifier stage may be used to drive the 80 watt booster amplifier, or the 6360 may be modulated to serve as a 15 watt transmitter.

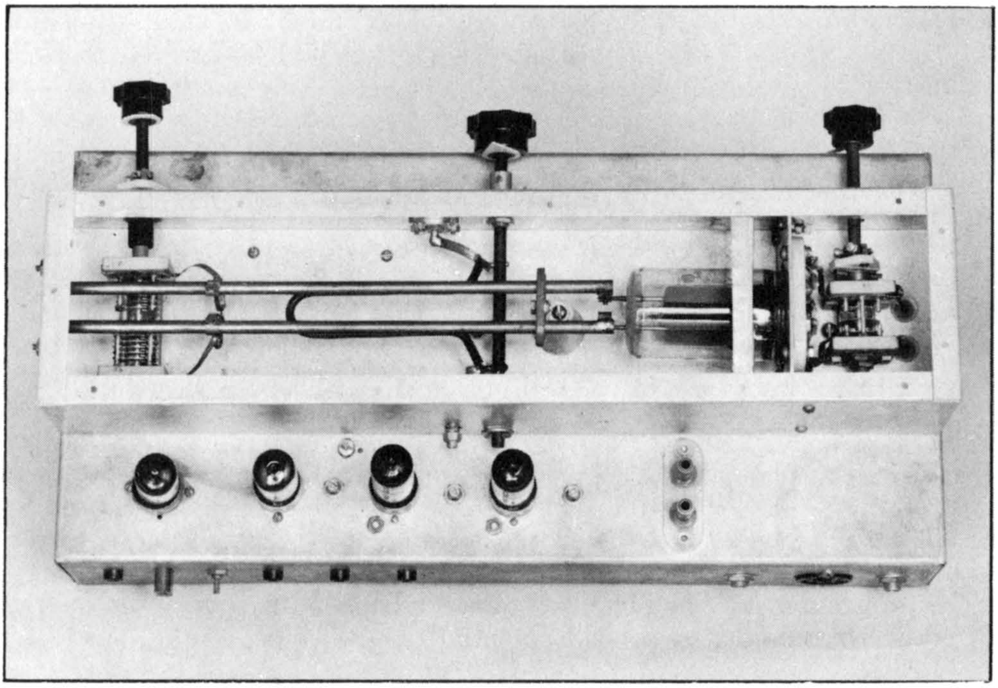


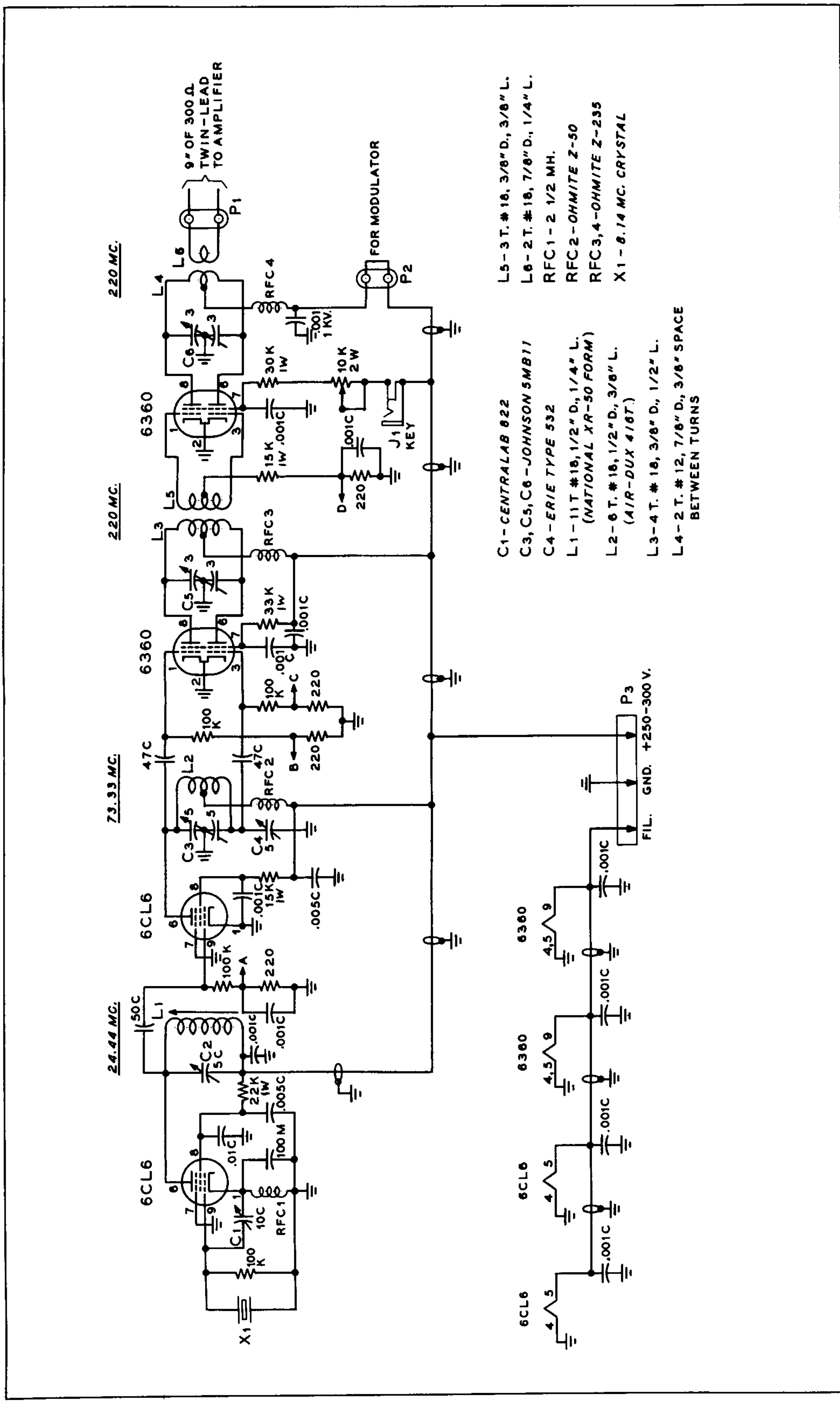
Fig. 15 220 mc transmitter uses 5894 double tetrode as amplifier in linear tank circuit. Complete shield is placed around the plate circuit of amplifier stage to reduce radiation loss from lines. Antenna coupling loop is at center.

The internal screening of the 6360 is sufficiently good to permit it to be used on the 220 mc band without neutralization, although a slight amount of interaction between the grid and plate circuits is in evidence.

Shown in Figure 17 is the schematic of the 80 watt amplifier, using a single 5894 tube. Loaded half-wave transmission lines are used for both the grid and plate resonant circuits of this stage. The grid line is resonated at the tube end by a split-stator variable capacitor (C7). A small amount of capacitive loading is present at the opposite end of the line in the form of the distributed capacity of the grid chokes. Excitation is coupled to the center of this resonant circuit through a short section of 300 ohm transmission line and a U-shaped loop (L7).

The plate circuit of the amplifier is enclosed in a separate compartment, shown in Figure 15. A shield is placed across the base of the 5894 tube to remove a slight degree of interaction between the grid and plate circuits. No neutralization of the tube is then required. The plate circuit of the amplifier is composed of a loaded half-wave section of transmission line. Capacitive loading is used at the far end of the line to shorten the line enough to allow it to fit inside the plate section of the amplifier box. A tuning capacitor is placed near the electrical center of the line, and a pickup loop (L10) is mounted at the center of the line. When operated at a plate potential of 450 volts, and a total plate current of 200 milliamperes (90 watts input) a measured output of 50 watts can be obtained from this amplifier.

Suitable meter shunts are incorporated in the grid circuits of the exciter and in the grid and plate circuits of the 5894 amplifier to permit circuit adjustment. In addition, a simple protective bias supply for the 5894 tube


24.44 MC.
73.33 MC.
220 MC.
220 MC.
6CL6
6CL6
6360
6360
9" OF 300 Ω TWIN-LEAD TO AMPLIFIER
FOR MODULATOR
J1 KEY
GND.
+250-300 V.
FIL.
C1-CENTRALAB 822
C3, C5, C6-JOHNSON 5MB11
C4-ERIE TYPE 532
L1-11 T #18, 1/2" D., 1/4" L. (NATIONAL XR-50 FORM)
L2-6 T. #18, 1/2" D., 3/8" L. (AIR-DUX 416T.)
L3-4 T. #18, 3/8" D., 1/2" L.
L4-2 T. #12, 7/8" D., 3/8" SPACE BETWEEN TURNS
L5-3 T. #18, 3/8" D., 3/8" L.
L6-2 T. #18, 7/8" D., 1/4" L.
RFC1-2 1/2 MH.
RFC2-OHMITE Z-50
RFC3,4-OHMITE Z-235
X1-8.14 MC. CRYSTAL

is used. Cathode keying of the power amplifier, or screen keying of the exciter may be used for c-w operation, if desired.

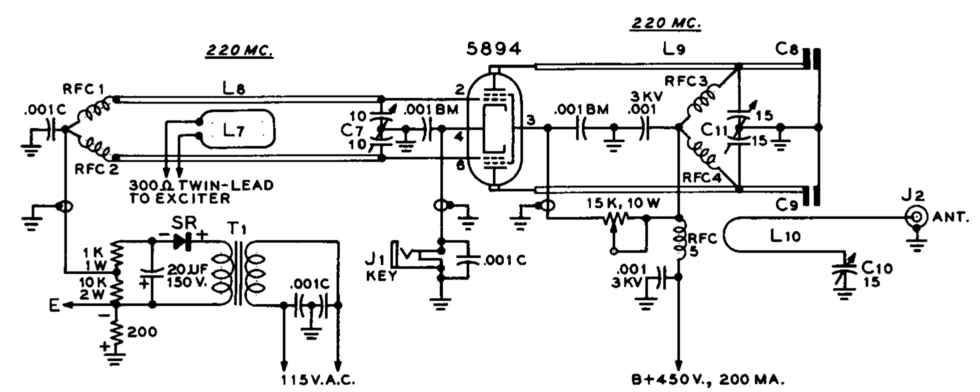
Exciter Construction

The complete transmitter is built upon an aluminum chassis measuring 8″ x 17″ x 3″ *(Bud AC-412)*. The exciter section is placed along the rear edge of this chassis, occupying a space 2½″ wide and 12″ long. The four tube sockets of the exciter stages are spaced 2¼″ apart. The 8.14 mc crystal socket, oscillator plate coil L1 and the grid current measuring jacks are placed in a line along the center of the rear lip of the chassis. The three midget butterfly tuning capacitors (C3, C5, and C6) are mounted on the chassis deck, spaced midway between the tube sockets. Balancing capacitor C4 is mounted to the chassis adjacent to the stator of C3. Plate coils L2, L3, and L4 are mounted directly to the stators of the butterfly capacitors. Grid coil L5 is supported directly from pins 1 and 3 of the 6360 amplifier socket.

The plate leads from the 6360 tube sockets to the butterfly tuning capacitors are made from short lengths of 3/16″ copper strap, and the various bypass capacitors, bias resistors, and r-f chokes in the exciter assembly mount between tube socket pins, or to phenolic tie-point strips mounted near the tube sockets.

The "hot" filament pin of each exciter tube socket is bypassed directly at the tube socket with a .001 ufd disc ceramic capacitor, and the interconnecting filament and plate leads use shielded wire. The outer braid of this wire is grounded at each end of the run.

The pickup link (L6) is attached to a two terminal assembly (P1). The 300 ohm balanced line coupling the exciter to the amplifier is also attached to this assembly. The line should be disconnected from the terminal assembly if it is desired to use the exciter without the amplifier.



L7 – 1-TURN LOOP # 12, 3″ LONG, 1″ WIDE.
L8 – TWO 3/16″ D. COPPER RODS, 11″ LONG, SPACED 3/4″ CENTER-TO-CENTER.
L9 – TWO 1/4″ D. COPPER RODS, 10 3/8″ LONG, SPACED 3/4″ CENTER-TO-CENTER.
L10 – 1-TURN LOOP # 12, 3 1/2″ LONG, 1″ WIDE.
SR – 50 MA. REPLACE. TYPE SELENIUM RECTIFIER.

C7 – *BUD LC-1664*
C8, C9 – *SEE TEXT*
C10 – *HAMMARLUND HF-15*
C11 – *BUD LC-1665*
RFC1–RFC5 – *OHMITE Z-235*
T1 – 125 V. AT 50 MA. *STANCOR PA-8421*

Fig. 17 Schematic of 220 mc 5894 amplifier stage. Safety bias supply shown.

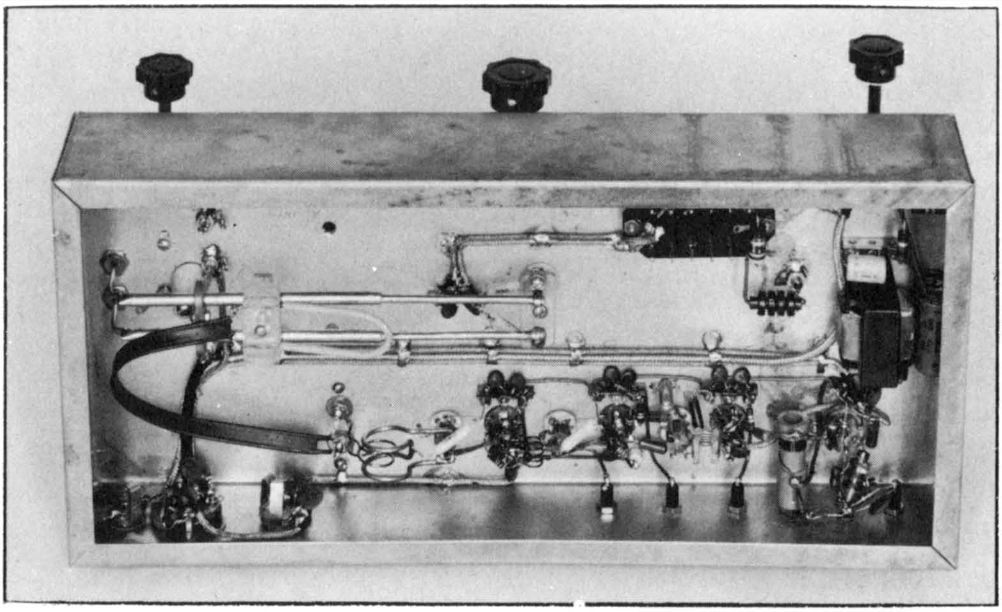


Fig. 18 Half wave grid lines of 5894 amplifier are shown in under-chassis view of 220 mc transmitter. Exciter stages are located along rear of the chassis. A short length of 300 ohm ribbon line couples grid lines to the plate circuit of 6360 driver stage. Safety bias supply is placed at upper right.

Exciter Adjustment

The exciter wiring should be checked, and the two 6CL6 tubes and an 8.14 mc crystal plugged in their respective sockets. 300 volts is applied to the exciter and the slug of coil L1 adjusted for crystal oscillation. Cathode capacitor C1 is set for optimum crystal feedback, evidenced by ease of starting of the oscillator. Grid current to the 6CL6 multiplier should run approximately 1 to 1.5 milliamperes, as measured at *point A*. The plate circuit of the second 6CL6 stage should now be resonated and balanced for equal grid current to each section of the 6360 multiplier. Start with C4 at minimum capacity and resonate C3. Slowly increase the capacity of C4, reresonating C3 with each change of C4. This process is repeated until equal grid current is obtained at *points B* and *C*. Grid current at each point should be at least 0.8 milliampere. The 6360 multiplier tank circuit, and the coupling between L3 and L5 are adjusted for an amplifier grid current of 3 milliamperes. The spacing between the turns of L5 should be adjusted for proper grid current when the plate circuit of the tripler stage is resonated. The 6360 amplifier may be loaded to a plate current of 75 milliamperes at a potential of 250 volts.

Amplifier Construction

The amplifier tube and plate circuit assembly is mounted above chassis in an aluminum box measuring 17″ x 4″ x 3″ *(Bud AC-432)*. The 5894 tube socket is mounted on a vertical plate located 3″ from the grid end of the box. Grid tuning capacitor C7 is located within the grid area of the box, and connections between the capacitor and the tube socket are made with short lengths of copper strap. The grid line passes through two polystyrene bushings into the under chassis area where it is supported by polystyrene strips mounted on two 1″ ceramic standoff insulators. Loop L7 is affixed

to a polystyrene sandwich which is bolted to the grid lines and may be slid back and forth to vary the coupling.

The plate lines are $10\frac{1}{4}''$ inches long, and are supported at the tube end by a polystyrene plate mounted in a vertical position atop a 1″ ceramic insulator. The opposite end of each line is soldered to a small copper tab measuring $\frac{3}{4}'' \times 1''$ in size. These tabs are separated from the end of the box-chassis by two thicknesses of Teflon insulation, forming capacitors C8 and C9. Teflon insulated bolts pass through the copper tabs and the box, and the capacity of each unit may be varied slightly by tightening the bolts. The purpose of these capacitors is to load the free end of the line so that is short enough to fit inside the shielded compartment. The plate tuning capacitor is mounted on the chassis, $1\frac{1}{2}''$ from the end of the box, and the stators are attached to a point approximately 3″ from the end of the line by means of two plate clips. Copper strap is employed for the leads to capacitor C11.

A tube shield plate is cut from thin aluminum to fit the inside of the chassis-box. A hole is cut in the center to pass the 5894 tube, and the shield is placed so that it lies in the same plane as the internal shield of the tube, providing excellent isolation between the input and output circuits.

Output link L10 is attached to a bakelite rod that passes through the walls of the plate compartment. The rod may be turned to vary the degree of coupling between the link and the plate tank circuit. Connections are made to the link by means of thin, flexible copper strap.

Amplifier Adjustment

The plate tank circuit should first be resonated to 220 mc with the 5894 in the socket. Capacitors C8 and C9 should be adjusted, and the taps to tuning capacitor C11 moved back and forth on the rods until resonance is established. Coupling between the exciter and the amplifier should now be set by the positioning of loops L6 and L7. Grid current should be adjusted to approximately 8 milliamperes with loop L7 positioned about the midpoint of the grid line. A dummy load is now attached to the antenna terminals and plate power applied to the amplifier. The coupling of L10 and the capacity of C10 should be adjusted to provide a loaded resonant plate current of 200 milliamperes to the 5894 tube. When excitation is removed, the action of the built-in bias supply will drop the plate current of the amplifier stage to a safe value.

420 Mc Driver-Amplifier

Shown in Figures 19 to 21 is a driver-amplifier for 420 mc operation. This unit is designed to be used with an existing 144 mc transmitter capable of supplying five or six watts excitation to the 420 mc unit. A 2 meter transmitter, such as the *Gonset Communicator* will serve as a satisfactory driver for 420 mc operation. By using existing 2 meter equipment, a minimum of circuit duplication is necessary for 420 mc activity.

It is only necessary to triple frequency to reach the 420 mc amateur band from the 2 meter region. In this case, a push-pull 6252 tetrode acts as a frequency tripler to 432 mc, and is inductively coupled to a push-pull 5894 class C 420 mc amplifier. Input to the 5894 stage is approximately 65 watts.

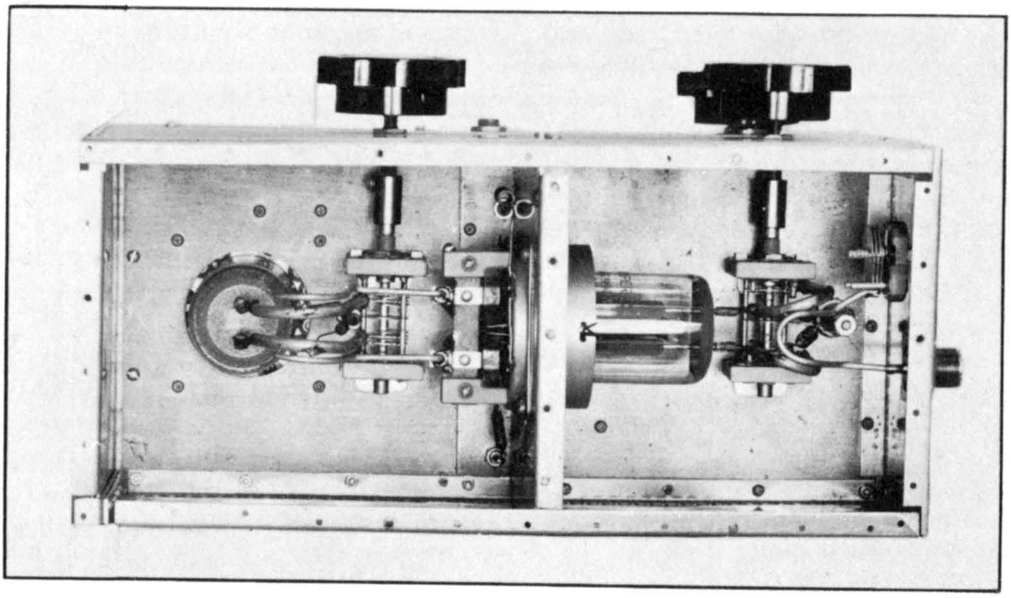


Fig. 19 Complete shielding is placed around 420 mc tuned circuits to reduce radiation loss. Top plate of enclosure is made of perforated aluminum. Holes are drilled around base of enclosure to enhance flow of air thru the box.

At this level, a power output of approximately 25 watts is obtained at the antenna terminals of the amplifier.

It is necessary to thoroughly shield the tripler and amplifier stage to reduce the effects of energy radiation from the tuned circuits of the transmitter. All power leads passing from these enclosures make use of high frequency feedthru type capacitors to reduce power lead radiation which can be very detrimental at these frequencies. Halfwave tuned line tank circuits are used in all 432 mc circuits to permit the greatest degree of circuit efficiency.

Transmitter Construction

The 420 mc driver-amplifier is built upon an aluminum chassis measuring 7″ x 14″ x 2″. Half-wave linear tank circuits are employed in the tripler and amplifier stages. These tuned circuits are enclosed in two compartments, visible in Figure 19. The compartments are 4½″ high, and have a row of ¼″holes drilled around the bottom to admit cooling air. The tops of the compartments are covered with perforated aluminum sheet. The socket of the 6252 tripler tube is a submounting unit, permitting the tube to project through the chassis. The 5894 socket is mounted vertically in the center of the partition separating the two compartments.

A box measuring 4½″ x 5″ is made of sheet aluminum, and is placed around the under-chassis grid circuit components of the 6252 tripler stage. A home-made split stator capacitor is used for C1 of Figure 20 but the unit specified in the parts list will work equally well. The capacitor is mounted parallel to the front of the amplifier, and is driven by a right angle coupling unit. The socket for the 6252 tripler has built-in mica bypass capacitors for the power pins. Additional bypassing and filtering is employed in the screen lead to this tube to reduce unwanted r-f currents and spurious couplings that might otherwise exist. Filament chokes are also

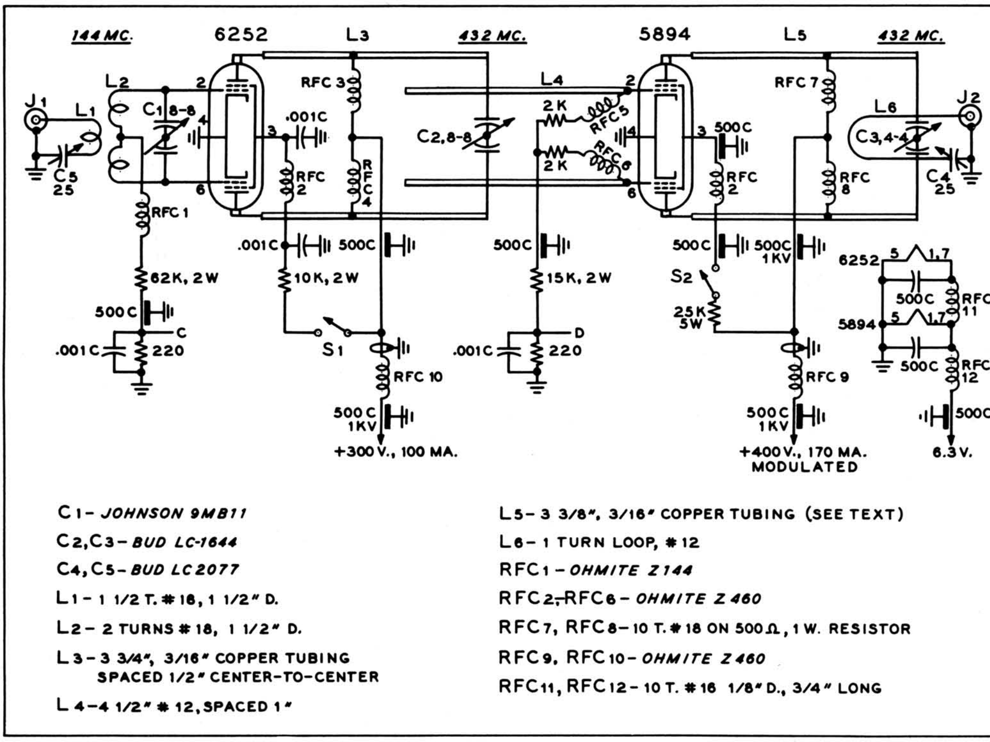


Fig. 20 420 mc operation is obtained by 144 mc tripler driving tetrode amplifier.

necessary in the leads to the 6252 and 5894, as shown in Figure 20.

The half-wave plate inductors of the two 432 mc tank circuits are made of lengths of 3/16″ silver plated copper tubing. The inductors are supported at the stator terminals of C2 and C3. The spacing between the rods is varied slightly until the tuned circuit resonates at 432 mc with the tuning capacitors set at approximately mid-scale. This test must be made with both tubes in their sockets, and the plate connections attached to the tubes. Short pieces of flexible copper braid or strap are used to make connection between the rods and the plate pins of the tubes. B-plus connection is made approximately 1¼″ inches from the plate end of the rods.

The grid inductor of the 5894 amplifier stage is made of two pieces of silver plated #12 wire. The wires extend horizontally from the tube socket

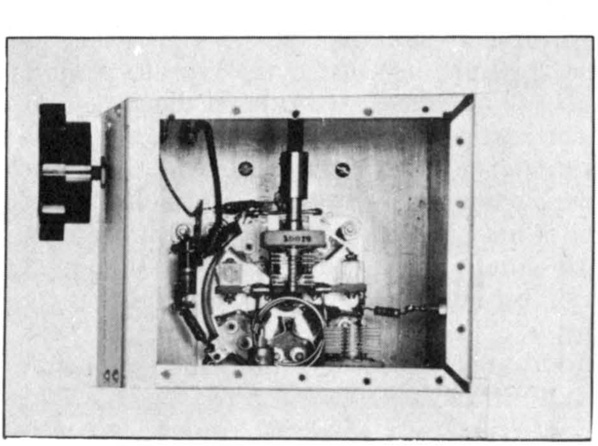


Fig. 21 Grid circuit of 420 mc tripler stage is enclosed to reduce radiation losses. Tripler may be driven by any 144 mc transmitter capable of 5 watts output.

until they reach the rods comprising L3, then they bend upwards and follow the contour of L3. RFC-5 and RFC-6 are mounted directly to socket pins 2 and 6 of the 5894 stage.

The rotors of C2 and C3 are left ungrounded, and these two capacitors are mounted about ¼″ above the metal chassis on short phenolic posts to improve the capacity to ground balance of the circuits.

Upon completion of the wiring, all circuits should be set to approximate frequency with the aid of a grid dip oscillator, and the wiring should be carefully checked for errors before plate voltage is applied to the unit.

Transmitter Adjustment

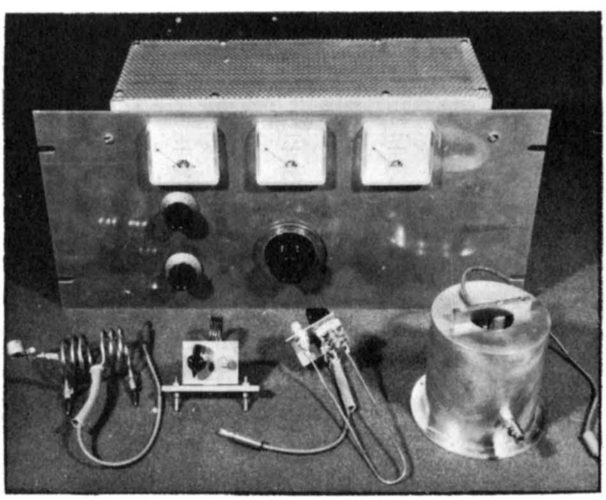
After all wiring has been checked, filament voltage may be applied to the tubes. Screen voltage should not be applied to the tubes, although it is permissible to apply plate voltage at the same time as the filament voltage. When the tuned circuits are near resonance, screen voltage may be applied to first the multiplier stage, and then to the amplifier.

The first step is to apply 2 meter excitation to the 6252 frequency multiplier. Coupling and loading adjustments should be made to produce a grid current of approximately 3.5 milliamperes in the tripler stage. This will drop to approximately 3 ma under load. Switch S1 is then closed, and C2 resonated at 432 mc. Coupling between L3 and L4 is adjusted for maximum grid current to the 5894 stage. Plate current of the 6252 is approximately 90-100 ma. It may be necessary to prune the length of the grid rods to achieve optimum interstage coupling. Grid current to the 5894 should run over 5 milliamperes, as measured at *point D*. The antenna is next connected to the amplifier stage, and S2 closed. C3 is tuned for resonance, and C4 and L6 adjusted for optimum antenna coupling. Under full input, grid current should run over 4 ma, and the plate current of the 5894 should be approximately 170 milliamperes.

4X-250B Amplifier for 50-420 Mc

This power amplifier is designed for the serious VHF amateur who wishes to use high power on the VHF bands. The amplifier is rated at an input of 300 watts, but may be run at 500 watts input on any band up to

Fig. 22 This tetrode amplifier is capable of operation on any band between 50 mc and 420 mc. Plug-in plate circuits of 4X250B tube are shown in foreground. Resonant "pot" tank is used for 420 mc.



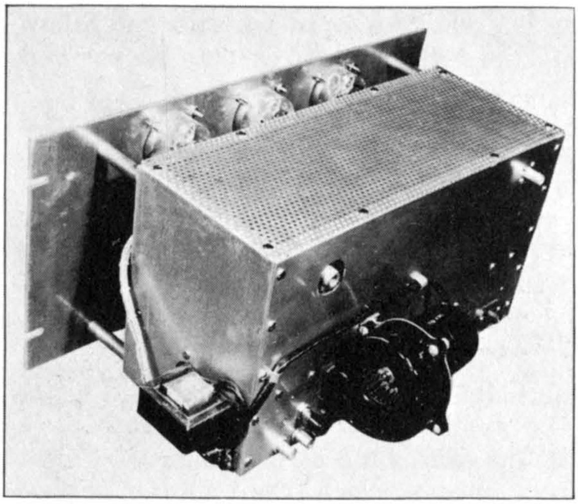
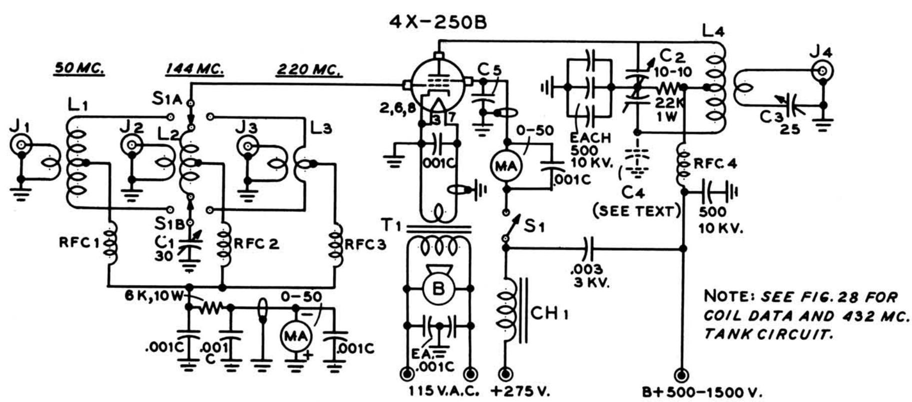


Fig. 23 4X250B is forced air cooled by blower motor mounted on rear of chassis. Air flows into closed grid compartment, passes through tube socket and escapes via perforated metal top of the amplifier cabinet.

220 mc. On the 420 mc band, power input is limited to 300 watts input. Present restrictions on that band, however, restrict the maximum power input to 50 watts. The amplifier, with the plug-in plate tank circuits is shown in Figures 22 and 23.

Circuit Description

The schematic of this multi-band amplifier is shown in Figure 24. A 4X-250B external anode tetrode is used in a conventional circuit that has



C1- *HAMMARLUND HF-30X*

C2- *MILLEN 12035* (35-35 μμF) WITH 2 CENTER ROTOR PLATES REMOVED.

C3- *JOHNSON 25J12*

C4- 1" X 1" COPPER SHEET, 3/4" FROM CHASSIS SUPPORTED BY END OF INDUCTOR.

C5- PART OF *EIMAC AIR SOCKET SK-610*

S1A, S1B- 2 WAFER CERAMIC SWITCH, *2P3T* SPACED 1" BETWEEN WAFERS. (MADE FROM *CENTRALAB 2543*)

RFC1- *OHMITE Z-50*

RFC2- *OHMITE Z-144*

RFC3- *OHMITE-Z-235*

RFC4- CHANGED WITH L4. USE PROPER *OHMITE* CHOKE FOR EACH BAND.

B- *RIPLEY* 115 V. BLOWER WITH *# 2 1/2-LR* IMPELLER.

T1- *STANCOR P-5014*

CH1- 16 H. AT 50 MA. *STANCOR C-1003*.

Fig. 24 Schematic of 4X250B VHF amplifier, showing "series tuned" grid circuit.

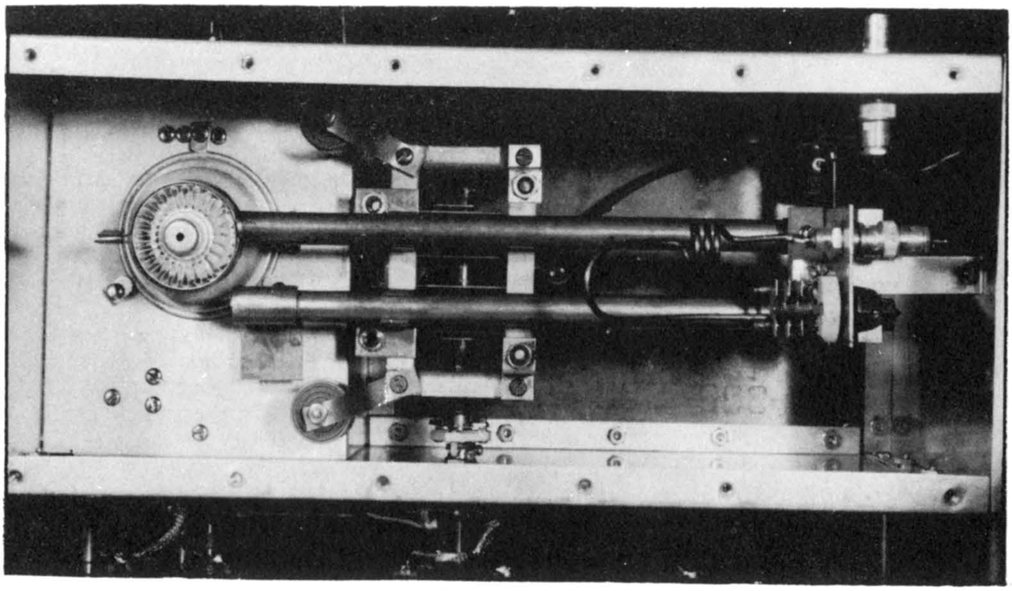


Fig. 25 Interior view of plate circuit compartment of 4X250B amplifier. The 144 mc tuned lines are in place. Lines are replaced by coil for 50 mc operation, and by resonant loaded cavity for operation at 420 mc. Cavity drops over plate of 4X250B tube, and is fastened to chassis by wing-nuts.

been modified for VHF use. The 4X-250B is used in conjunction with the *Eimac SK-610* air socket having built-in VHF capacitors for the screen terminals. The properties of this socket permit maximum stability to be achieved in this amplifier, and no neutralizing or parasitic suppression circuits are required.

The amplifier grid circuit employs a band switching turret covering 50, 144, and 220 mc. A series tuned configuration is used to obtain the best L-C ratio at the higher frequencies. The plate tank circuit used at 50 mc is a standard coil-capacitor arrangement. The 144 mc and 220 mc bands use a capacity loaded parallel line circuit which is balanced to ground. For 420 mc operation, a tuned cavity is substituted for the parallel line tank circuit. Shown in Figure 28, this cavity is a capacity loaded ¼ wavelength section of coaxial line which fits over the top of the 4X-250B tube (Figure 27). The cavity bolts directly to the chassis over the tube. Contact to the plate of the 4X-250B is made by a circular strip of finger stock, insulated from the inner conductor of the cavity by a length of teflon which acts as a plate blocking capacitor. Plate voltage is shunt fed to the tube through an r-f choke and a plate connector, made from a length of phosphor bronze strap.

Resonance is established by a disc-type tuning capacitor mounted in the side of the cavity facing the front panel. The capacitor is adjusted from the panel by means of a shaft extension assembly visible in Figure 25.

For operation above the six meter band, coil L4 is removed from the coil jacks mounted on the left side of C2, and the 144-220 mc plate assembly is plugged into the fuse clips mounted on C2. The banana plug of the plate assembly fits into a matching jack mounted on a ceramic insulator at the end of the plate compartment. This is only a mechanical support, as the plate lead from the movable shorting bar plugs into an insulated jack mounted on the front wall of the amplifier box. The antenna coupling

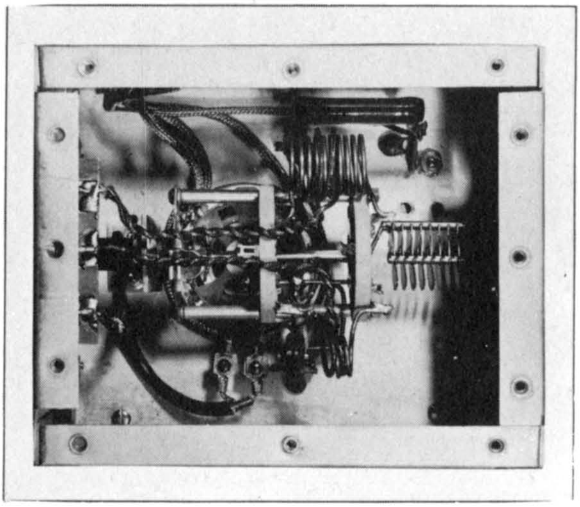


Fig. 26 Interior of grid compartment of 4X250B amplifier. Grid coils are placed between wafers of ceramic switch. The 220 mc grid inductance is beneath the switch, and not visible. Extra switch contacts permit installation of 28 mc coil, if desired.

loops for 144 and 220 mc plug into a receptacle assembly mounted to the far end of the amplifier box, as shown in Figure 25. The 50 mc antenna coil plugs into banana jacks mounted on the far side of tuning capacitor C2. A short length of coaxial line completes the antenna circuit from the pickup loop assembly to the coaxial receptacle mounted on the rear wall of the amplifier box.

Connection to the plate of the 4X-250B is made by means of a thin copper strap soldered to the end of the 144-220 mc line which encircles the anode of the tube.

Amplifier Assembly

The amplifier is built within an aluminum box measuring 6″ x 8″ x 17″. The 4X-250B socket is mounted upon a small enclosure at one end of the box. The grid circuit components are placed within this enclosure. When the bottom plate is placed upon the box, the grid compartment is effectively sealed from the plate circuit area. The 115-volt blower motor is mounted on the rear of the box with the port projecting into the grid circuit area. The air passes through the tube socket and out the top of the amplifier plate compartment. Grid circuit wiring is done with 1/8″ copper strap.

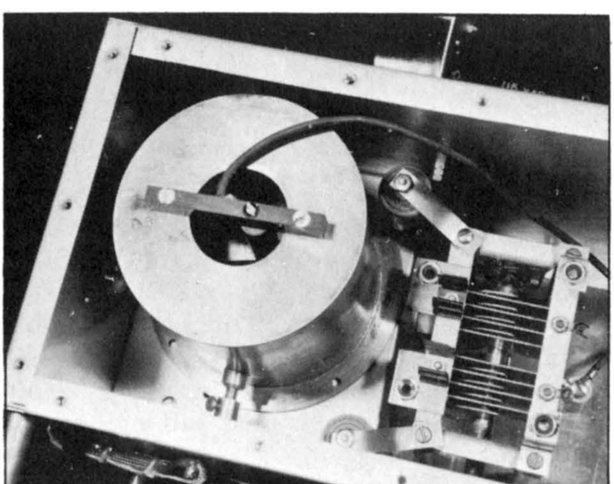
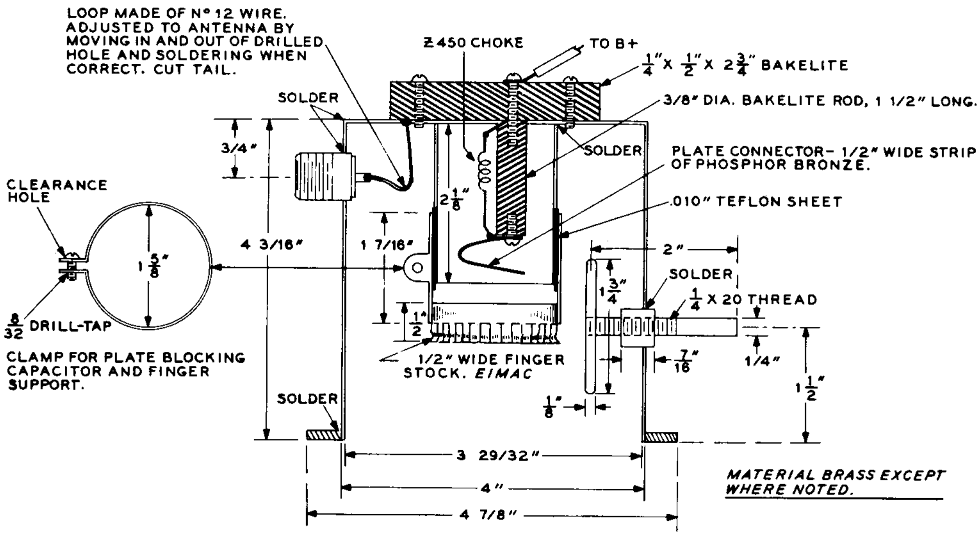


Fig. 27 420 mc plate circuit cavity shown at left. Assembly drawing of cavity is shown in Figure 28. Outer shell of cavity is at ground potential, as plate of tube is shunt-fed.



COIL TABLE

L1 – 10 T. #14 E., 5/8" D., 1" LONG. LINK 2 TURNS HOOKUP WIRE WOUND ABOUT CENTER OF L1

L2 – 2 T. #14 E., 3/8" D., 3/8" LONG. LINK–SAME AS L1.

L3 – 1" #14 E. BETWEEN SWITCH CONTACTS (SEE TEXT)

L4 – 50 MC. – 6 T. #10, 1 1/2" I.D., 2 1/2" L., LINK 4 TURNS #10.
144 MC. – 10 1/4" L., 1/2" TUBING SPACED 1 1/4" CENTER-TO-CENTER. LINK 3 T. #10, 1/2" D. IN SERIES WITH LOOP OF #10, 2 1/2" L., 1 1/8" WIDE.
220 MC. – SAME AS 144 MC. WITH SHORT AND COUPLING LOOP MOVED TO 6" FROM TUBE. LINK–LOOP 5 1/4" L., 1 1/4" W.

Fig. 28 Assembly drawing of 420 mc plate circuit cavity and plate clamp.

Placement of grid circuit components may be seen in Figure 26. The six and two meter coils are visible, but the 220 mc coil is mounted beneath the switch. This coil is a length of #14 wire running between the switch contacts. Length of the wire is varied slightly to establish circuit resonance. The link for L3 is a square loop of wire placed near the grid coil.

Plate tuning capacitor C2 is bypassed at both ends and at the middle with TV-type ceramic capacitors to place it at ground potential on all bands. In addition, the two middle rotor plates of C2 are removed to provide a bearing surface on the rotor shaft for a phosphor bronze wiper which is connected to the center bypass capacitor. This contact is necessary on 220 mc for proper grounding of the tuning capacitor.

It is necessary to silver solder all connections in the plate tank circuit, especially the shorting strap on the 144-220 mc tuned line because of the high circulating r-f current. The amplifier operates at an overall efficiency of about 58%, and a certain proportion of the lost power goes into heating the amplifier plate circuit components. A good silver plating job will help to reduce plate circuit losses to a degree.

Amplifier Operation

Amplifier operation is very straightforward. No neutralization or parasitic suppression is necessary. For 420 mc operation, the 4X-250B can double from 216 mc, or triple from 144 mc. The grid current required for proper operation is 16 to 18 milliamperes. The efficiency of the stage at 420 mc can be improved by increasing the grid bias resistor to 15K, 20 watts. Normal plate voltage is 1500, although the voltage should be dropped for 420 mc operation to stay within the power limitations on that band.

A High Power Grounded-Grid Amplifier for 144 Mc

The legal maximum power is desired for scatter signal transmission or moon reflection work at 144 mc. For greatest reliability, it is necessary for the equipment to run at this level with a considerable safety margin. A transmitter running on the "ragged edge," or in danger of breaking down at any moment is worse than useless. Shown in Figures 29 to 33 is a grounded-grid 144 mc power amplifier of proven reliability. Under test it has run 1800 watts input into a suitable dummy load for hours at a time (3000 volts at 600 ma). It has been keyed at an input of 3000 watts (4000 volts at 720 ma) in commercial service for extended periods of time with excellent results. At an input of 1000 watts, the ample reserve of this unit will permit many trouble-free hours of operation.

Because the grounded-grid circuit is inherently degenerative, the amplifier is extremely stable, requiring no neutralization or parasitic suppression. The overall power gain is such that input to the driver should be approximately ¼ the amplifier input when the latter is operated at 3000 volts. A pair of 826's running at slightly more than 250 watts input are capable of sufficient drive for inputs greatly exceeding 1-kilowatt. The g-g amplifier uses a pair of 592 triode tubes, and operates at a plate efficiency of about 70%. To this output must be added the portion of grid driving power passed through the amplifier from the exciter. Actual r-f output at 1-kilowatt input is thus in excess of 800 watts.

Amplifier Circuit

The circuit of the g-g amplifier is shown in Figure 31. The grids of the 592 triodes are bypassed to ground and provided with a safety bias of -150 volts. The 592 tubes have a split filament which is parallel connected. Filament voltage is fed to each tube through a high current r-f choke, and the circuit between the tubes is completed through the filament inductor, L1.

To achieve optimum symmetry, a balanced input circuit is used. Two taps are placed on L1, and r-f excitation is directly applied to the filament inductor. If it is desired to employ coaxial line to couple the amplifier to the exciter, a balun may be placed across the input terminals, as shown in

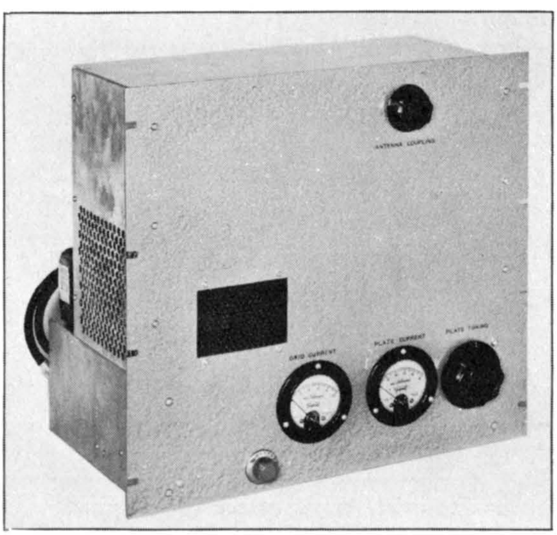


Fig. 29 Kilowatt amplifier employs grounded grid circuit with two Eimac 592 triode tubes. Exceptional circuit stability achieved with this configuration. Amplifier will operate with plate voltage in excess of 4000 with no parasitics.

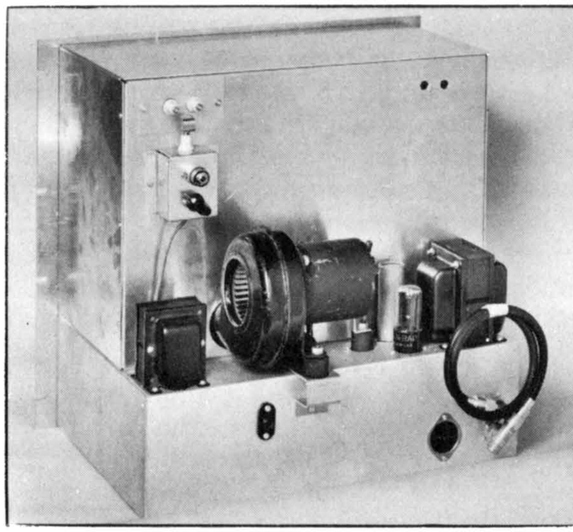
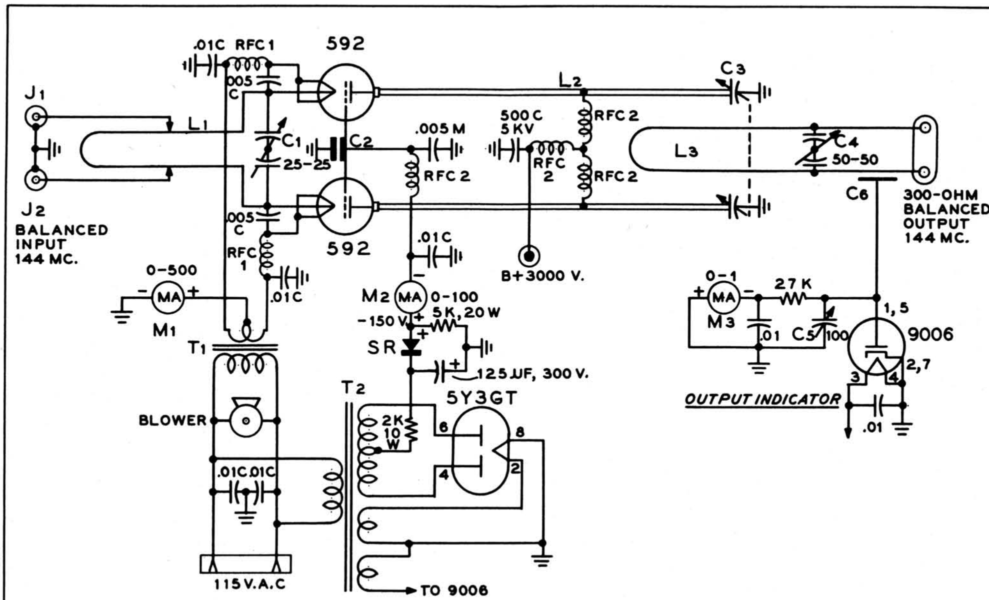


Fig. 30 Blower motor to cool tubes is mounted on rear of amplifier chassis. Push-pull input circuit is fed from unbalanced coaxial line by means of flexible coax balun placed between input terminals.

Figure 30. The termination of the center conductor of the coaxial line should not be returned to ground in the exciter, as it will short out the cathode meter, M1, in the amplifier. Series tuning of the pickup link in the exciter should therefore be used. A high-C filament tank provides maximum



C1-25-25 UUF, *BUD LC-1666*

C2-BYPASS CAPACITOR PLATE, 9" X 4 1/2" INSULATED FROM CHASSIS BY *"MYLAR"* SHEET. .005" THICK (APPROX. 1000 UUF)

C3-TWO *JENNINGS VAC-40* VARIABLE VACUUM CAPACITORS, GANGED BY FOUR GEARS-*BOSTON G-464*

C4-50-50 UUF, *BUD CE-2041*

C6-1/2" X 1" METAL TAB PLACED NEAR OUTPUT TERMINALS.

T1-*MERIT P-3146*, 10 V., 10 A.

T2-*STANCOR PC-8403*, 500 V. CT., 40 MA., 6 V., 5 V.

L1-HAIRPIN, #8, 2" LONG X 5/16" WIDE TAP 1/2 WAY UP FROM BOTTOM.

L2-1/16" X 1 1/4" ALUMINUM STRAPS, 19 1/4" LONG, SPACE 3 1/2" C-C, PLUS FLEXIBLE PLATE LEADS 2" LONG, EACH. HORIZONTAL PORTION OF STRAPS IS 13" LONG.

RFC1-36 T. #14 E. 3/8" D., CLOSE WOUND.

RFC2-*OHMITE Z-144*

INPUT BALUN-27" RG-8/U, INCLUDING PLUGS.

SR-100 MA. SELENIUM RECTIFIER.

J1, J2-SO-239 RECEPTACLES.

Fig. 31 Schematic of kilowatt grounded-grid 144 mc amplifier stage.

drive to the grounded grid amplifier, and C1 should tune to 144 mc when it is almost completely meshed.

A half-wave line is used in the plate circuit of the amplifier. The far end of the line is tuned by means of two ganged variable vacuum capacitors. A balanced link provides 300 ohm output termination for a balanced line, and a simple vacuum tube voltmeter using a 9006 diode is employed as an r-f output indicator.

Bias for the 592 tubes is supplied from a simple supply incorporated in the amplifier. A selenium rectifier is used to disconnect the supply when the rectified bias voltage exceeds the voltage of the supply.

Amplifier Construction

The amplifier is built upon an aluminum chassis measuring 13″x17″x5″. The plate circuit compartment measures 12″x9½″x17″. The two 592 tubes are located at the left end of the enclosure, with the sockets mounted 2¾″ below the chassis on a small plate. The twin grid pins of the tubes are positioned about ½″ above the chassis. Grid bypass capacitor (C2) is made of a sheet of aluminum measuring 4½″x8″, and is separated from the chassis by a sheet of teflon. The grid pins of each tube are attached to this plate.

The two *Jennings* 40 uufd variable vacuum capacitors (C3) are placed at the right of the chassis in a vertical position. They are panel driven by a bevel gear train visible in Figure 33. The plate inductors (L2) run from the plate caps of the tubes to the stators of the tuning capacitors.

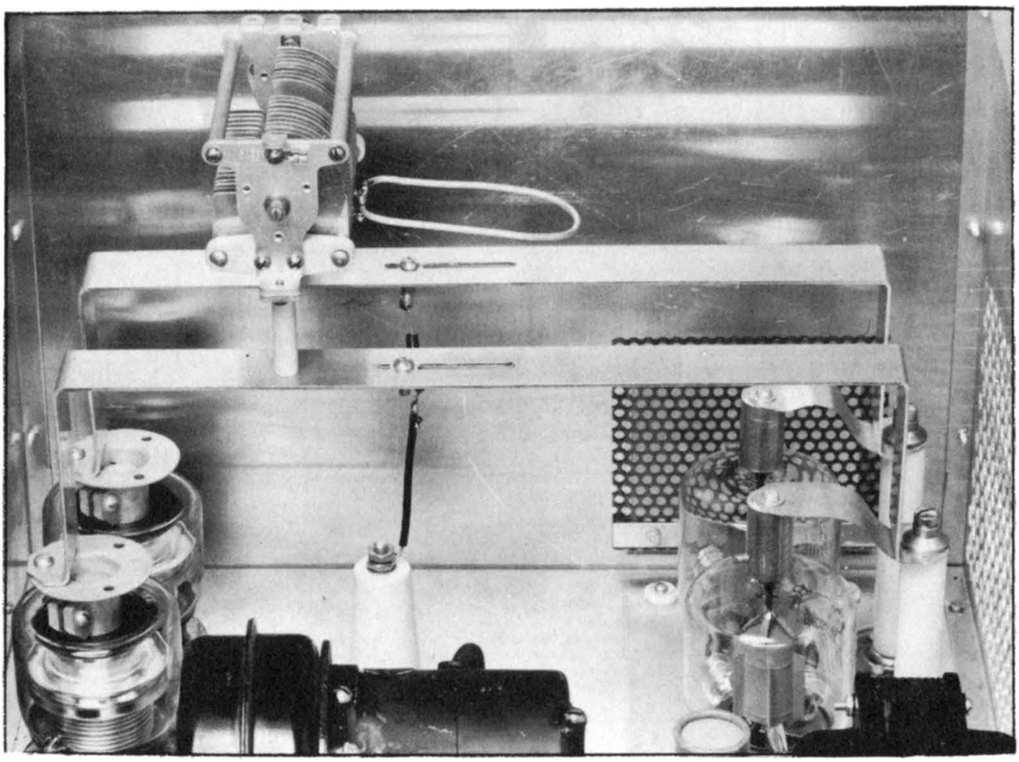


Fig. 32 View of interior of plate circuit enclosure of grounded grid amplifier. Plate lines are made of aluminum strap and are an electrical half wave in length. The far end of lines is tuned by two ganged vacuum variable capacitors. Antenna tuning capacitor is mounted atop lines.

Positioning of the blower, the output indicator and the bias supply can be seen in Figure 30. Also shown is the input balun mounted on the coaxial input receptacles.

The filament tuning capacitor (C1) is mounted to the filament socket plate by a small angle bracket. All filament wiring should be made of short, heavy leads. The input leads from the coaxial plugs to L1 are made of short lengths of 52 ohm coaxial line (RG-58/U). The outer shield is grounded at each end of the lines.

Amplifier Adjustment

Grid and plate circuits should be adjusted to resonance with the aid of a grid dip oscillator. Excitation should be applied to the amplifier, and the taps on L1 adjusted for maximum grid current (50 ma) with minimum input to the exciter. The amplifier should be connected to a dummy load, and low plate voltage applied. Cathode meter M2 reads grid current, and rectified r-f plate current in addition to the true value of plate current, and tuning is best done with the aid of the vacuum tube voltmeter. Plate tuning should be adjusted for maximum output with minimum color of the plates of the tubes. Excitation should be increased, and plate voltage should be raised until the desired input is run. Be careful to guard against exceeding the grid dissipation of the tubes. Excitation should always be decreased when plate voltage is removed from the tubes. For complete tuning procedure of a grounded grid amplifier, the reader is referred to the *Radio Handbook*, published by Editors and Engineers, Summerland, Calif.

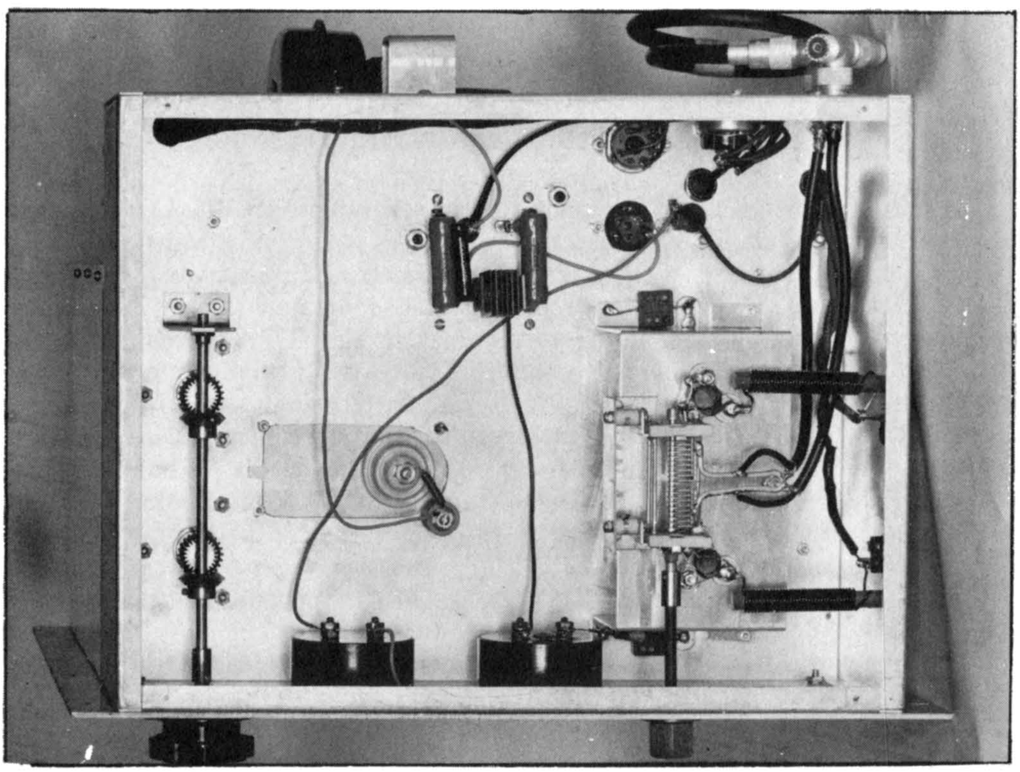


Fig. 33 Under-chassis view of grounded grid amplifier. Filament chokes are visible near tubes, and input "hairpin" is attached to side of split stator tuning capacitor. Excitation taps may be varied to adjust drive.

CHAPTER XII

VHF Test Equipment

There are several pieces of test equipment that are vital to the VHF enthusiast. Two of the most important items are the Noise Generator and the Standing Wave Meter (SWR Bridge). Two practical versions of these instruments are described in this chapter. No well equipped VHF station should be without one of each of these vital pieces of equipment. In addition, an accurately calibrated grid dip oscillator and a VHF frequency standard are essential tools for the construction and calibration of VHF equipment.

Noise Generators and VHF Receivers

In the VHF and UHF range the ultimate sensitivity of a receiver or converter is limited by its noise figure. The atmospheric noise in this frequency range is quite low, and the major sources of noise are the Johnson noise generated by the antenna, and the noise created by the passage of electrons in the r-f tubes of the receiver. Any signal has to compete with this collection of noise in the receiver. The quantitative concept of this noise is termed the *noise figure* (N/F) of the receiver. It is important to reduce the noise figure as low as possible. If the receiver noise figure is reduced from 15 db (a not uncommon figure) to 5 db, it has the same effect as raising by ten times the power output of the station being received.

It is common practice to employ a *noise generator* to determine the noise figure of a VHF receiver or converter. The noise generator produces a "white" noise that covers many megacycles of the spectrum. The amount of this special noise required to boost the output of the receiver over that value obtained when the noise generator is inoperative is a measure of the noise figure.

There are a number of different types of noise generators. They are all alike, however, in the fact that their available noise power is proportional to a d.c. current. The simplest noise generator is a *silicon* crystal diode, having a d.c. current passing through it in the reverse (high resistance) direction. A more precise instrument may be made from a temperature-

limited diode, which acts as a constant current noise generator by virtue of the shot effect of the electron stream within the tube. For general use, the crystal noise generator is simple, inexpensive and easy to build and is to be preferred to the more precise and expensive diode noise generator.

The Crystal Noise Generator

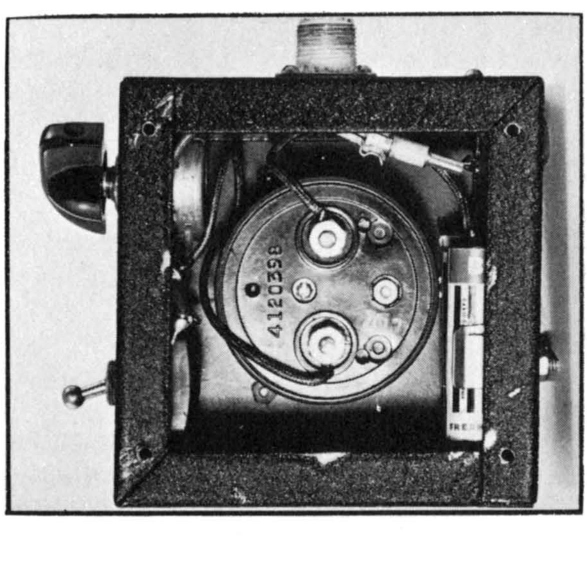
A simple silicon crystal diode noise generator suitable for use with 52 ohm, or 72 ohm termination is shown in Figure 1. A surplus 1N23 crystal is used, held in position with a fuse clip and a pin taken from an octal socket. The circuit of the simple instrument is shown in Figure 2. This generator does not give noise figure in actual db. It is a comparison device only, and merely tells the user whether the adjustments he makes on his receiver or converter are in the right direction of lower noise figure. With the particular crystal and components used in the noise generator shown, it was found that one milliampere of reverse crystal current (indicated on the noise generator meter) is equal to approximately 10 db of noise when compared with a laboratory noise generator.

The generator is built in a small aluminum box measuring 2″ x 4″ x 4″ *(Bud AU-1083)*. The two-inch meter is mounted on one face of the box, and the coaxial receptacle on one side. The silicon crystal and R2 are mounted to the center pin of the receptacle with as short leads as possible. The .001 ufd ceramic capacitor is grounded at one of the flange bolts of the receptacle. The potentiometer and battery switch are placed on the lower side of the box, and the 3-volt battery is mounted in a metal clip on the top wall of the box, as shown in Figure 2. Use of this instrument will be discussed later in this chapter.

The Vacuum Noise Generator

The shot effect of a diode tube makes a very effective noise source. The amount of noise may be easily controlled by varying the filament current of the diode. The best diode tubes employ pure tungsten or thoriated tungsten filaments. One of the best tubes to use is the *Sylvania* type 5722.

Fig. 1 A simple noise generator may be made from a silicon diode having a d.c. current passed through it in the reverse (high resistance) direction. Shown at right is interior view of such a unit, built in a 2″ x 4″ x 4″ box. Schematic is shown in Fig. 2.



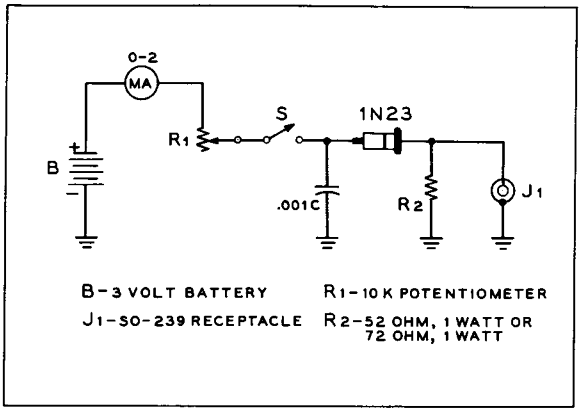


Fig. 2 Schematic of simple crystal diode noise generator. Noise output is a function of diode current, as measured on self-contained milliammeter. Unit is powered by three flash-light cells.

The transmitting-type 24G also performs in a satisfactory manner.

A practical diode noise generator must be well shielded, and the power leads should have sufficient shielding and filtering to prevent pickup of undesired signals, and to prevent unwanted coupling to the input circuit of the receiver under test. Some form of filament voltage control should be incorporated in the device, and also a meter to read the plate current of the noise diode.

The noise generator is terminated in a load resistor whose ohmic value is equal to the impedance of the transmission line with which the receiver is to be used. This resistor is mounted directly across the coaxial receptacle of the noise generator, using the shortest possible leads.

Generator Construction

The diode noise generator and power supply are housed in a cabinet measuring 7″ x 10″ x 8″ *(Bud C-993)*. The unit is built upon an aluminum chassis 7″ x 9″ x 2″ *(Bud AC-406)*. The powerstat controlling the filament voltage, the 0-5 d.c. milliammeter and the decibel meter are mounted at the top of the front panel, as seen in Figure 3. The 5722 socket is mounted to the chassis directly behind the SO-239 coaxial receptacle, and the 52 ohm 1-watt carbon terminating resistor is mounted between the center pin of the receptacle and the grounding ring of the diode socket. The power wiring is not critical, it is only necessary to employ the shortest possible leads for the filament bypass capacitors on the socket of the 5722. Using an 0-5 d.c. milliammeter, the maximum obtainable noise level is 7 db. A 0-15 milliammeter will permit a noise level of 11.7 db. When this value of termination is used the noise figure is ten times the logarithm of the diode plate current in milliamperes. For other values of resistive termination the noise figure is:

$$(1) \quad NF_{(DB)} = 10 \log_{10} 20 \times IR$$

where I is the diode plate current in *amperes*, and R is the terminating resistance in ohms.

In use, the noise generator is coupled to the input of the receiver or converter under test and the diode filament voltage is increased until the noise output power of the receiver is double that read with the diode in-

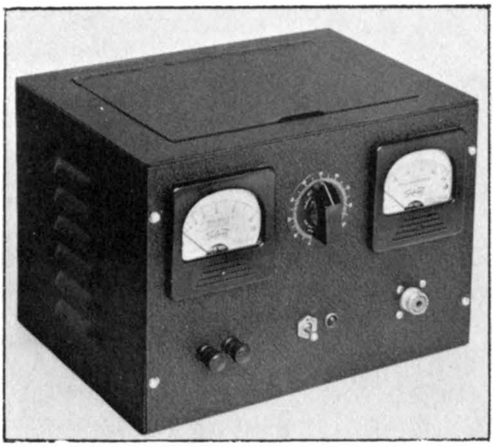


Fig. 3 Front view of vacuum tube noise generator.

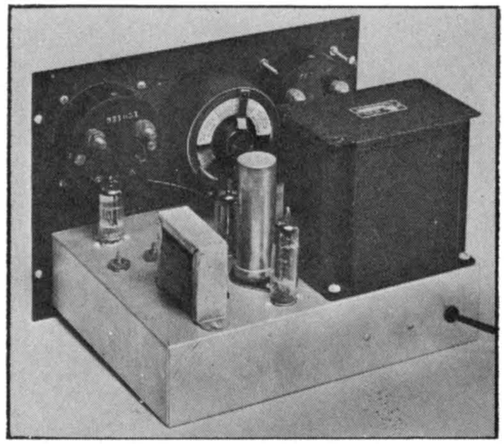


Fig. 4 Rear view of vacuum tube noise generator.

operative. The value of diode current under these conditions, when substituted in formula 1 will indicate the noise figure in decibels of the equipment under test.

Use of the Noise Generator

The test procedure is essentially the same whether a silicon or a vacuum diode noise generator is used. The generator should be connected to the antenna terminals of the receiver under test by means of a short length of coaxial cable. For accurate results in the VHF region, it is important that coaxial fittings be used at each end of the line. With the noise generator off, and the r-f gain control of the receiver open, the audio gain control should be adjusted until a reading is obtained on the db output meter of the noise generator. If desired, a separate vacuum tube voltmeter or rectifier type meter may be employed. A reference level should be established, and

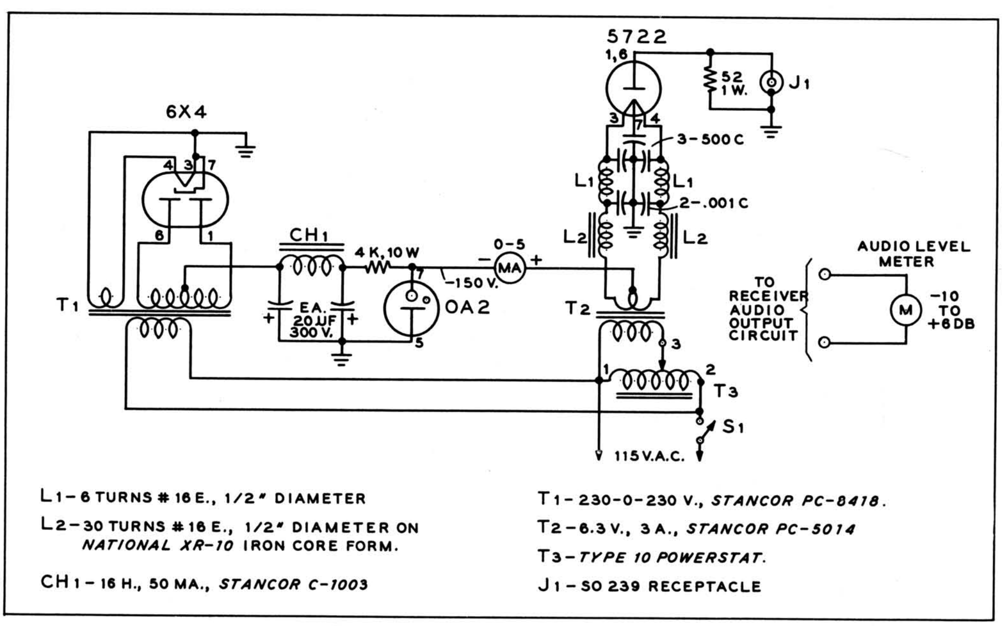


Fig. 5 Schematic of vacuum tube noise generator using 5722 noise-diode tube.

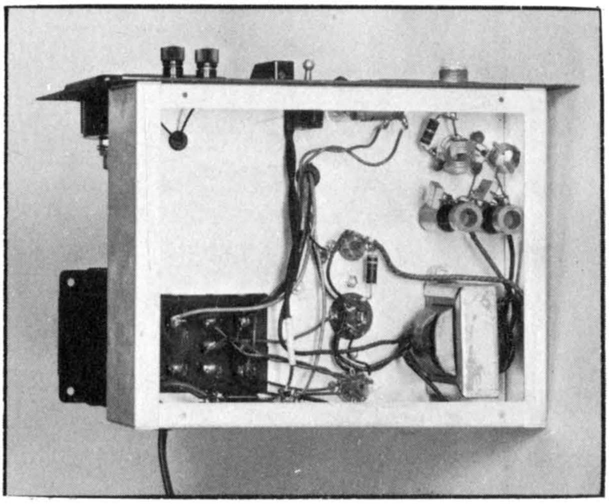


Fig. 6 Under-chassis view of 5722 noise generator. Noise diode tube socket is placed next to coaxial output plug, and connections to plug are made with heavy copper strap.

the audio gain control should not be touched thereafter.

The next step is to turn on the noise generator and adjust the filament voltage control (or the crystal diode current) until the audio power level of the receiver has doubled. This is an increase of 3 db on the audio level meter, or an increase of 41% on an a.c. voltmeter. The value of noise generator current at this point is used for computation of the noise figure of the receiver.

In the case of the diode noise generator, no accurate measurement of the true noise value can be obtained. The object therefore is to make various circuit adjustments which will result in the same percentage increase in the reading of the output meter, but with lower and lower readings on the noise generator meter. The same procedure can be used with the vacuum diode instrument. As the amount of noise generated in the receiver is reduced by adjusting the tuned circuits and the interstage coupling, a smaller and smaller amount of external noise injection is required to activate the output meter on the receiver in the same proportion.

Adjustments should be made to the antenna input tap, or the antenna coil, the tuning of the input circuit and subsequent circuits, and the interstage coupling. In general best noise figure will be obtained when the input circuit is slightly overcoupled.

When the vacuum diode noise generator is used the measurements are independent of receiver bandwidth, and comparisons may be made directly between one receiver and another providing they both have the same input impedance. In addition the noise generator may be employed to neutralize cascode circuits, or in any test where a low level r-f source is required.

For operation above 150 mc, it is often necessary to resonate the output circuit of the noise generator to the frequency of measurement. This may be done with a small parallel tuned circuit, mounted directly at the terminals of the diode tube. The circuit is tuned for maximum noise output of the generator.

Interpretation of Noise Readings

The noise generator is primarily intended for receiver alignment, and for determination of the lowest value of noise figure. Caution should be exerted, however, when a specific value of noise figure is derived from the

measurements. The technique of determining the *actual value* of noise figure is tricky, at best. Unfortunately, almost all errors in measurement technique tend to make the noise figure seem better than it really is. Care must be taken to eliminate spurious signal paths, and the second detector of the receiver should be examined for linearity to make sure that it does not provide a misleading and nonlinear reading. For most accurate results the output of the receiver should be measured by a vacuum tube voltmeter placed across the last i-f stage, rather than by direct audio measurements.

A Simple Standing Wave Indicator for Balanced Lines

Shown in Figure 7 is a simple "twin lamp" type of standing wave indicator, suitable for use in the VHF region. It may be clipped to a balanced transmission line by means of the spring-type binding posts. The enamel on the line should be cleaned away at the point of contact. The large loop lies in the plane of the line, parallel to the wires and about 1/4" away. When the line is energized, one or both pairs of lamps will light. If the line is properly matched, the lamps towards the transmitter will be lit, and the lamps nearest the antenna will be dark. As the value of SWR increases, the antenna lamps will glow brighter. All antenna adjustments should aim for greatest brightness of the transmitter lamps, and least illumination of the antenna lamps.

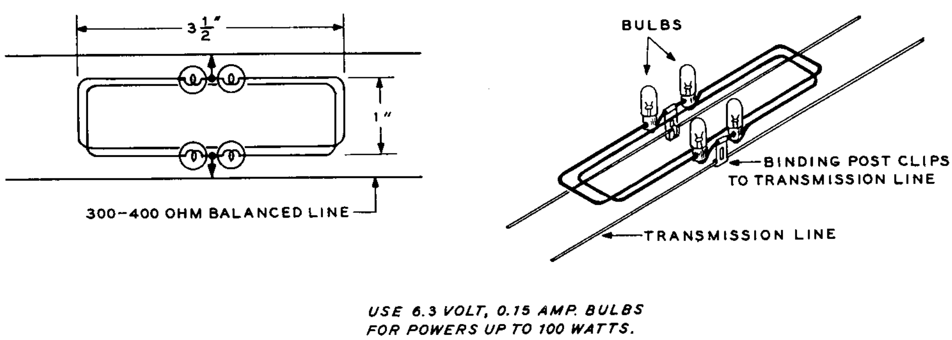


Fig. 7 A simple SWR meter for open wire lines is made from loop of wire and four flashlight bulbs. Loop is attached to transmission lines by means of two clips. It is positioned above the line, and parallel to it. When line has unity SWR, lamps nearest transmitter will light, and others remain out.

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